MARCONI INSTRUMENTS LIMITED

UK

Head Office
Longacres
ST. ALBANS
Hertfordshire AL4 0JN
Telephone: (0727) 59292
Telex: 23350
Fax: (0727) 57481 (Gp. 3)

Service Division
The Airport
LUTON
Bedfordshire LU2 9NS
Telephone: (0582) 33866
Telex: 825248

Microwave Products Division
Gunnels Wood Road
STEVENAGE
Hertfordshire SG1 2AU
Telephone: (0438) 312311
Telex: 82159

FRANCE

MARCONI INSTRUMENTS
3 Avenue du Maréchal Devaux
91550 PARAY-VIEILLE-POSTE
Telephone: (1) 687 36 25
Telex: 203882 F

WEST GERMANY

MARCONI MESSTECHNIK GmbH
Postfach 153
Landsberger Strasse 65
8034 GERMERING
Telephone: 089-84 50 85
Telex: 5212642

USA

MARCONI INSTRUMENTS
Division of Marconi Electronics Inc.
100 Stonehurst Court
NORTHVALE
New Jersey 07647
Telephone: (201) 767-7250
TWX: 710-991-9752

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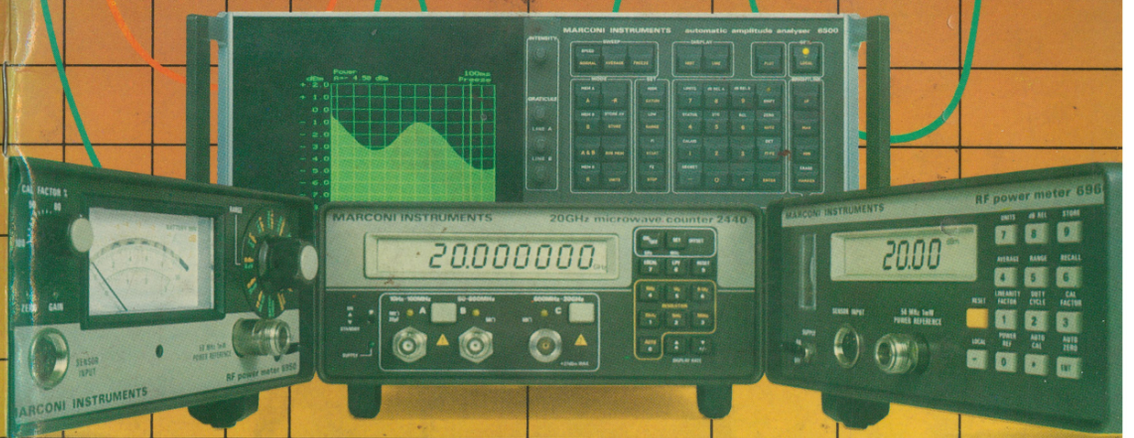
Microwave Datamate



MICROWAVE DATAMATE

A collection of useful reference data for microwave and RF engineers.

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GENERAL

Units & Conversion Factors

Angular Measure:
1° (deg) = right angle/90 = 0·01745329 rad
1 radian (rad) = 57°.29578 = *3437′·747 = 206264″·8*
1 milliradian (m rad) = 3′26″·26

1 *steradian (sr) = 3282·8 sq.deg.*

Length:
1 micron (μm) = 1 micrometre = 10^{-6}m = 0·0001cm = 10,000 Å
1 ångstrom (Å) = 10^{-10}m = 0·0001μm
1 mil = 0·001 inch (in)
1 in = 2·540 000 cm 1 foot (ft) = 30·48 cm 1 yard (yd) = 0·9144m
1 metre (m) = 39·37008 in = 3·280840 ft = 1·093613 yd
1 mile = 1760 yd = 5280 ft = 8 furlongs = 80 chains = 320 rods
= 1·609344 km = 0·8689762 nautical mile (international)
1 km = 0·6213712 mile 5 miles = 8·046720 km
1 nautical mile (international) = 1852 m
1 fathom = 6 ft = 1·8288 m

Velocity & Angular Velocity:
1 m.p.h. = 1·6093 km/hr = 88 ft/min = 1·4667 ft/sec = 44·7040 cm/sec
1 ft/sec = 30·48 cm/sec = 0·6818 m.p.h. = 1·097 km/hr
1 UK knot = 1 UK nautical mile/hr = 1·151515 mile/hr = 1·688889 ft/sec
= 1·85318 km/hr = 51·48 cm/sec = 1·00064 knot (international)
1 km/hr = 0·6214 m.p.h. = 0·9113 ft/sec = 27·78 cm/sec
1 velo = 1 cm/sec = 0·0328 ft/sec
Mach 1 = velocity of sound at a given height
Mach 1 at ground level (15°C & 14.7 lb/sq.in) = 1116·1 ft/sec
= 1224·7 km/hr = 760·97 m.p.h. = 660·85 UK knot = 661·27 knots (international)
Unit angular velocity = 1 rad/sec = 9·549 rev/min = 0·1592 rev/sec
= 57°·3/sec = 2·06265 x 10^5 deg/hr
1°/hr = 1′/min = 1″/sec = 4·848 x 10^{-6} rad/sec

Acceleration:
1 ft/sec^2 = 0·3048 in/sec^2 1in/sec^2 = 3·281 ft/sec^2
1 Gal (or 1 celo) = 1 cm/sec^2 = 0·0328 ft/sec^2
Standard g = 9·80665 m/sec^2 = 32·1740 ft/sec^2

Area:
1 sq. in. = 6·4516 sq. cm. 1 sq. cm. = 0·1550 sq. in.
1 sq. ft. = 0·0929 sq. m.
1 sq. yd. = 9 sq. ft. = 0·8361 sq. m.
1 sq. m. = 10·764 sq. ft. = 1·196 sq. yd.
1 acre = 10 sq. chains = 43560 sq. ft. = 4840 (= 69·57^2) sq. yd.
1 sq. mile = 640 acres = 258·999 hectares = 2·58999 sq. km.
1 hectare = 10,000 sq. m. = 2·471 acres
1 barn (b) = 10^{-24} sq. cm.
1 sq. Å = 10^{-16} sq. cm. = 100 Mb.
Area of 1st Bohr orbit of hydrogen atom = πa_0^2 = 0·87974 x 10^{-16} sq. cm. = 0·88 sq. Å

Volume:
1 cu. in. = 16·387 cu. cm. (ml) 1 cu. m. = 35·31 cu. ft. = 1·308 cu. yd.
1 cu. yd. = 0·7646 cu. m.
1 UK gallon (= 1 gal.) = 4·54609 l = 277·42 cu. in.
1 US gallon = 3·785411 l = 231 cu. in.
1 litre (l) = 1 cu. dm.
1 cu. ft. = 6·229 UK gal = 7·481 US gallons

Pressure:
1 Standard Atmosphere (0°C) = 760 mm. Hg. = 29·921 in Hg. = 14·696 lb wt./sq. in.
= 1·01325 x 10^6 dyn/sq. cm. = 1·01325 Bar = 1033·23 g wt/sq. cm. = 101325 N/sq. m.
= 33·90 ft. water 4°C.
1 Torr = 1 mm Hg. = 1·359 g wt/sq. cm. = 1333 dyn/sq. cm. = 1·333 m bar
1 g wt/sq. cm. = 980·7 dyn/sq. cm. = 98·07 N/sq. m.
1 lb wt./sq. in. = 70·307 g wt/sq. cm. = 68·95 m bar = 51·72 Torr.
1 millibar (m bar) = 1000 dyn/sq. cm. = 100 N/sq. m. = 0·0145 lb wt./sq. m. = 0·7502 Torr.

Mass:
1 lb (avoirdupois) = 0·45359237 kg. = 16 oz = 7000 grain = 1·2153 lb Troy or Apothecary.
1 grain (all systems) (gr) = 64·7989 mg.
1 kg. = 2·20462 lb (a.v.)
1 metric tonne = 1000 kg = 10^6 (gram) g.
1 metric carat = 0·2 g.
1 oz = 28·3495 g.
1 slug = 32·174 lb = 14·5939 kg.
1 lb (US) = 0·45359243 kg.

Force:
1 lb wt. = 4·4482 x 10^5 dyn 1 g wt. = 980·665 dyn.
1 poundal (pdl) = 1·3825 x 10^4 dyn
10^6 dyne (dyn) = 1·0197 kg. wt. = 2·2481 lb. wt.
1 Newton (N) = 10^5 dyn.

Density:
1 lb/cu. in. = 27·680 g/cu. cm.
1 g/cu. cm. = 0·03613 lb/cu. in.
1 lb/cu. ft. = 16·0185 kg/cu. m.
1 kg/cu. m. = 0·06243 lb/cu. ft.

Water density (4°C) = 0·999972 g/cu. cm. = 62·43 lb/cu. ft.
Mercury density (0°C) = 13·5951 g/cu. cm.

Temperature:
°C = 5 (F−32)/9 °F = (9C/5) + 32
°F = Fahrenheit °C = Celcius (formerly Centigrade)
K = degrees Kelvin = °C + 273·15
°R = Rankine °R = °F + 459·67

Energy:
1 ft. lb wt. (= loosely, 1 ft. lb.) = 32·16 ft. pdl. = 192 in. oz wt. = 13825 cm. g wt.
= 1·356 x 10^7 dyn cm. (= erg.) = 1·356 N.m. (or m N).
1 oz wt. in. = 72·01 g wt. cm.
1 N.m. = 1 Joule (J) = 10^7 erg. = 0·7376 lb wt. ft.
1 British Thermal Unit (Btu) = 1055 J = 252 c = 778 lb wt. ft.
1 Board of Trade Unit = 1 kW hr. = 3600 x 10^3 J = 2·655 x 10^6 lb wt. ft. = 3412 Btu.
1 Therm = 10^5 Btu
1 calorie 15°C (c) = 4·1855 J 1 calorie (international) = 4·1868 J
1 large calorie (C) = 1000c.

Moment of Inertia:
1 kg m^2 = 10^7g $cm.^2$ = 23·73 lb ft^2 = 0·7376 slug ft^2.
1 slug ft^2 = 32·174 lb $ft.^2$ = 13·558 x 10^6 g cm^2 = 1·3558 kg m^2.

Power:
1 watt (W) = 10^7 erg/sec. = 1 J/sec.
1 h.p. = 745·7W = 2544 Btu/hr = 42·4 Btu/min = 550 "ft lb"/sec = 178·2 c/sec.
1 metric h.p. = 735·499 W.

General:
π = 3·141592654 2π = 6·283185307 e = 2·718281829
Earth rotation rate w.r.t. fixed stars = 15°·041/hr = 7·292116 x 10^{-5} rad/sec.
Earth rotation rate w.r.t. sun = 15°/hr. = ¼°/min.
Mean radius of earth = 6·37103 x 10^6 m
Equatorial radius of earth = 3963·207 miles = a
Polar radius of earth = 3949·919 miles = c.
Ellipticity = 1/298·25 = (a − c)/a.
Eccentricity = 0·08182 = $(a^2 - c^2)^{1/2}/a$.

Metric Units

Exa	E	10^{18}	Deci	d	10^{-1}
Peta	P	10^{15}	Centi	c	10^{-2}
Tera	T	10^{12}	Milli	m	10^{-3}
Giga	G	10^9	Micro	μ	10^{-6}
Mega	M	10^6	Nano	n	10^{-9}
Kilo	k	10^3	Pico	p	10^{-12}
Hecto	h	10^2	Femto	f	10^{-15}
Deca	da	10	Atto	a	10^{-18}

Selected Mathematical Formulae

Trigonometrical Functions

$\sin (A + B) = \sin A \cos B + \cos A \sin B$
$\cos (A + B) = \cos A \cos B - \sin A \sin B$
$\sin (A - B) = \sin A \cos B - \cos A \sin B$
$\cos (A - B) = \cos A \cos B + \sin A \sin B$
$\sin 2A = 2 \sin A \cos A$
$\cos 2A = \cos^2 A - \sin^2 A = 1 - 2 \sin^2 A = 2 \cos^2 A - 1$

$\sin^2 A + \cos^2 A = 1$
$1 + \tan^2 A = \sec^2 A$
$1 + \cot^2 A = \operatorname{cosec}^2 A$

Hyperbolic Functions

$\sinh x = \frac{1}{2}(e^x - e^{-x})$
$\cosh x = \frac{1}{2}(e^x + e^{-x})$
$\tanh x = \dfrac{e^x - e^{-x}}{e^x + e^{-x}}$.

Series

Taylor's

$$f(x) = f(a) + (x-a)f'(a) + \frac{(x-a)^2}{2!}f''(a) + \frac{(x-a)^3}{3!}f'''(a) + \ldots$$

Maclaurin's

$$f(x) = f(0) + xf'(0) + \frac{x^2}{2!}f''(0) + \frac{x^3}{3!}f'''(0) + \ldots$$

Binomial

$$(1+x)^n = 1 + nx + \frac{n(n-1)}{2!}x^2 + \frac{n(n-1)(n-2)}{3!}x^3 + \ldots \quad (x < 1)$$

$$(1-x)^n = 1 - nx + \frac{n(n-1)}{2!}x^2 - \frac{n(n-1)(n-2)}{3!}x^3 + \ldots$$

Logarithmic

$$\log (1+x) = x - \frac{x^2}{2} + \frac{x^3}{3} - \frac{x^4}{4} + \ldots \quad (x < 1).$$

$$\log (1-x) = -x - \frac{x^2}{2} - \frac{x^3}{3} - \frac{x^4}{4} - \ldots$$

Exponential

$$e^x = 1 + x + \frac{x^2}{2!} + \frac{x^3}{3!} + \frac{x^4}{4!} + \ldots$$

Differentials

$d(ax) = a\,dx$
$d(uv) = u\,dv + v\,du$
$d\dfrac{(u)}{(v)} = \dfrac{v\,du - u\,dv}{v^2}$
$dx^n = nx^{n-1}\,dx$
$de^{ax} = ae^{ax}\,dx$
$da^x = a^x \log a\,dx$
$d(\sin x) = \cos x\,dx$
$d(\cos x) = -\sin x\,dx$
$d(\tan x) = \sec^2 x\,dx$
$d(\cot x) = -\operatorname{cosec}^2 x\,dx$
$d(\sec x) = \tan x \sec x\,dx$
$d(\operatorname{cosec} x) = -\cot x \operatorname{cosec} x\,dx$

Integrals

1. $\int x^n\,dx = \dfrac{x^{n+1}}{n+1} \quad (n \neq -1)$
2. $\int \dfrac{1}{x}\,dx = \log x$
3. $\int e^{ax}\,dx = \dfrac{e^{ax}}{a}$
4. $\int \log x\,dx = x \log x - x$
5. $\int u \dfrac{dv}{dx}\,dx = uv - \int v \dfrac{du}{dx}\,dx$
6. $\int \sin x\,dx = -\cos x$
7. $\int \cos x\,dx = \sin x$
8. $\int \operatorname{sh} x\,dx = \operatorname{ch} x$
9. $\int \operatorname{ch} x\,dx = \operatorname{sh} x$
10. $\int \dfrac{1}{a^2 + x^2}\,dx = \dfrac{1}{a}\tan^{-1}\dfrac{x}{a}$
11. $\int \dfrac{1}{\sqrt{a^2 - x^2}}\,dx = \sin^{-1}\dfrac{x}{a}$
12. $\int \dfrac{1}{\sqrt{(x^2 + a^2)}}\,dx = \operatorname{sh}^{-1}\dfrac{x}{a} = \log (x + \sqrt{a^2 + x^2})$
13. $\int \dfrac{1}{\sqrt{x^2 - a^2}}\,dx = \operatorname{ch}^{-1}\dfrac{x}{a} = \log (x + \sqrt{x^2 - a^2})$
14. $\int \dfrac{f'(x)}{f(x)}\,dx = \log f(x)$
15. $\int \tan x\,dx = -\log (\cos x)$
16. $\int \cot x\,dx = \log (\sin x)$
17. $\int \operatorname{cosec} x\,dx = \log (\tan \frac{1}{2}x)$
18. $\int \sec x\,dx = \log\left\{\tan\left(\dfrac{\pi}{4} + \frac{1}{2}x\right)\right\} = \log (\sec x + \tan x)$
19. $\int \dfrac{dx}{a + bx} = \dfrac{1}{b}\log (a + bx)$

FREQUENCY DESIGNATIONS

Telecommunications

Satellite Communications

Frequency Range (GHz)	Designation	Comments
2·5 – 2·69	Broadcast	Arabsat 2·6/6 GHz
3·7 – 4·2	Major Telecoms	Intelsat IV, V (Down) Telecom I & II
		4 – 17 GHz (Up/Down)
4·4 – 4·9	European Satellite	Symphonie (Down)
5·7 – 8·4	European Satellite	Symphonie (Up/Down)
5·925 – 6·4	Major Telecoms	Intelsat IV, V (Up)
10·7 – 11·7	Telecoms	OTS (MKI) & ECS 11·49 – 11·7 GHz (Down) ECS 11·45 – 11·78 GHz (Down)
		Intelsat IV, V (Down)
11·7 – 12·7	Broadcast	TV-Sat 11·7 – 12·7 GHz (Down)
14 – 15·35	Telecoms	OTS 14 – 14·5 GHz (Up)
		Intelsat IV, V (Up)
17	Broadcast	TV-Sat 17·3 – 18·1 GHz (Up)
17·3 – 19·7		New Satcoms
19 – 21·4		
22·5 – 23·6	Broadcast	
27 – 27·5		
39·5 – 40·5		
40·5 – 42·5	Broadcast	

European Telecommunication Satellites

ECS – European Communication Satellite (Launch 1983)
MARECS – Maritime Telecommunication Satellite
OTS – Orbital Test Satellite
DBS – Direct Broadcast Satellite
TV-Sat, DBS – Direct Broadcast Satellite (Launch 1983)
Telecom I & II – French Commercial Digital Telecom Satellite
USA Intelsat IV, V – Major Telecommunication Satellites over Atlantic and Indian Oceans for Worldwide Network

Terrestrial Links

Frequency Range (GHz)	Designation	Comments
1·4 – 1·53		**ITU Allocation Region 1**
1·7 – 2·7	2 GHz Band	60 – 960 ch FDM Systems
3·4 – 4·2	4 GHz Band (Lower)	1800 ch FDM Systems (BT) Future – Digital
4·4 – 5·0	4 GHz Band (Upper)	
5·85 – 6·4	6 GHz Band (Lower)	High Capacity 2700 ch FDM Systems (BT) Future – Digital 140 MBIT/S (BT)
6·4 – 7·0	6 GHz Band (Upper)	New TV/4 Voice ch FDM System (BT)
7·25 – 7·75	7 GHz Band	Low Capacity 300 ch FDM Systems
8·2 – 8·5	8 GHz Band	960 ch FDM Systems
9·3 – 10·68		
10·7 – 11·7	11 GHz Band	New Digital 140 MBIT/S 4 Phase Systems (BT)
11·7 – 12·5	14 GHz Band	Radio Relays (FDM) mainly TV and Studio Transmission over short range
12·75 – 13·25		
14·4 – 14·5		
14·5 – 15·5		
17·7 – 21·2	17 GHz Band	New Digital Radio Relay Systems
21·2 – 29·5		
31 – 31·3		Short Range Systems
36 – 40·5		
42·5 – 43·5		
47·2 – 50·2		

12 GHz Satellite Broadcasting Bands

Table shows Channel numbers, frequencies and country assignment.

Channel	Country	Frequency MHz	Channel	Country	Frequency MHz
1	F	11727·48	21	B	12111·08
2	D	11746·66	22	CH	12130·26
3	LUX	11765·84	23	NED	12149·44
4	GB	11785·02	24	I	12168·62
5	F	11804·20	25	B	12187·80
6	D	11823·38	26	CH	12206·98
7	LUX	11842·56	27	NED	12226·16
8	GB	11861·74	28	I	12245·34
9	F	11880·92	29	B	12264·52
10	D	11900·10	30	CH	12283·70
11	LUX	11919·28	31	NED	12302·88
12	GB	11938·46	32	I	12322·06
13	F	11957·64	33	B	12341·24
14	D	11976·82	34	CH	12360·42
15	LUX	11996·00	35	NED	12379·60
16	GB	12015·18	36	I	12398·78
17	F	12034·36	37	B	12417·96
18	D	12053·54	38	CH	12437·14
19	LUX	12072·72	39	NED	12456·32
20	GB	12091·90	40	I	12475·50

Note: Even No. Channels are left circular polarized and odd channels are right circular polarized.

Band Designation, Old and New

BAND DESIGNATIONS OLD	BAND DESIGNATIONS NEW	FREQUENCY GHz	WAVELENGTH
HF 3-30MHz	A 0-250MHz	0.1	3m
VHF 30-300MHz		0.15	2m
		0.2	1.5m
	B 250-500MHz	0.3	100cm
UHF 300-1000MHz		0.5	60cm
	C 500-1000MHz	0.75	40cm
L 1-2GHz	D 1-2GHz	1.0	30cm
		1.5	20cm
S 2-4GHz	E 2-3GHz	2	15cm
	F 3-4GHz	3	10cm
C 4-8GHz	G 4-6GHz	4	7.5cm
	H 6-8GHz	5	6cm
X 8-12GHz	I 8-10GHz	6	5cm
	J 10-20GHz	8	3.75cm
J Ku 12-18GHz		10	3cm
K 18-27GHz		15	2cm
	K 20-40GHz	20	1.5cm
Q Ka 27-40GHz		30	10mm
		40	7.5mm
40-60GHz	L 40-60GHz	50	6mm
		60	5mm
O E 60-90GHz	M 60-100GHz	75	4mm
		100	3mm

WG PARAMETERS

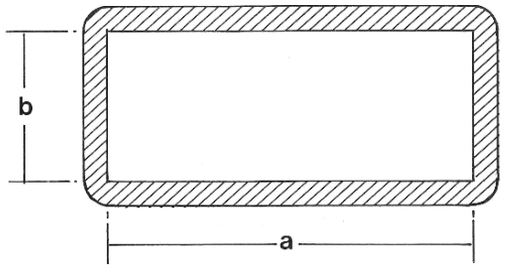
Rigid Rectangular Waveguide Data

Frequency Range (GHz) For TE_{10} Mode	Cut-off Freq. (GHz) For TE_{10} Mode	Theoretical CW Power Rating Lowest To Highest Freq. (MW)	Theoretical Attenuation dB/30m	Material Alloy	JAN RG ( )/U	WG Designation			Band Designations			Internal Dimensions	
						53-IEC	RCSC	IEA WR( )	UK	USA	New	(Inches)	(mm approx)
0·32-0·49	0·256	153·0-212·0	0·051-0·031	Alum		R3	WG00	2300			B	23·0 x 11·5	584·0 x 292·0
0·35-0·53	0·281	120·0-173·0	0·054-0·034	Alum		R4	WG0	2100			B, C	21·0 x 10·5	533·0 x 267·0
0·41-0·625	0·328	93·4-131·9	0·056-0·038	Alum	201	R5	WG1	1800			B, C	18·0 x 9·0	457·0 x 229·0
0·49-0·75	0·393	67·6-93·3	0·069-0·050	Alum	202	R6	WG2	1500			C	15·0 x 7·5	381·0 x 191·0
0·64-0·96	0·513	35·0-53·8	0·128-0·075	Alum	203	R8	WG3	1150	P		C	11·5 x 5·75	292·0 x 146·0
0·75-1·12	0·605	27·0-38·5	0·137-0·095	Alum	204	R9	WG4	975			C, D	9·75 x 4·875	248·0 x 124·0
0·96-1·45	0·766	17·2-24·1	0·201-0·136	Alum	205	R12	WG5	770			D	7·7 x 3·85	196·0 x 98·0
1·12-1·70	0·908	11·9-17·2	0·317-0·212	Brass	69	R14	WG6	650	L	L	D	6·5 x 3·25	165·0 x 83·0
			0·269-0·178	Alum	103								
1·45-2·20	1·157	7·5-10·7				R18	WG7	510			D, E	5·1 x 2·55	131·0 x 65·0
1·70-2·60	1·372	5·2-7·5	0·588-0·385	Brass	104	R22	WG8	430		LS, R	E	4·3 x 2·15	109·0 x 55·0
			0·501-0·330	Alum	105								
2·20-3·30	1·736	3·1-4·5	0·877-0·572	Brass	112	R26	WG9A	340			E, F	3·4 x 1·7	86·0 x 43·0
			0·751-0·492	Alum	113								
2·60-3·95	2·078	2·2-3·2	1·102-0·752	Brass	48	R32	WG10	284	S	S	E, F	2·84 x 1·34	72·0 x 34·0
			0·940-0·641	Alum	75								
3·30-4·90	2·577	1·6-2·2				R40	WG11A	229			F, G	2·29 x 1·145	59·0 x 29·0
3·95-5·85	3·152	1·4-2·0	2·08-1·44	Brass	49	R48	WG12	187	C	C	F, G	1·872 x 0·872	48·0 x 22·0
			1·77-1·12	Alum	95								
4·90-7·05	3·711	0·79-1·0				R58	WG13	159			G, H	1·59 x 0·795	40·0 x 20·0
5·85-8·20	4·301	0·56-0·71	2·87-2·30	Brass	50	R70	WG14	137			H	1·372 x 0·622	35·0 x 16·0
			2·45-1·94	Alum	106								

Rigid Rectangular Waveguide Data

Frequency Range (GHz) For TE_{10} Mode	Cut-off Freq. (GHz) For TE_{10} Mode	Theoretical CW Power Rating Lowest To Highest Freq. (MW)	Theoretical Attenuation dB/30m	Material Alloy	JAN RG ( )/U	WG Designations			Band Designations			Internal Dimensions	
						53-IEC	RCSC	IEA WR( )	UK	USA	New	(Inches)	(mm approx)
7·05-10·0	5·259	0·35-0·46	4·12-3·21	Brass	51	R84	WG15	112			I	1·122 x 0·497	29·0 x 13·0
			3·50-2·74	Alum	68								
8·20-12·40	6·557	0·20-0·29	6·45-4·48	Brass	52	R100	WG16	90	X	X	I, J	0·9 x 0·4	23·0 x 10·0
			5·49-3·83	Alum	67								
10·00-15·00	7·868	0·17-0·23				R120	WG17	75			J	0·75 x 0·375	19·0 x 9·5
12·40-18·00	9·486	0·12-0·16	9·51-8·31	Brass	91	R140	WG18	62	J	Ku	J	0·622 x 0·311	16·0 x 7·9
			—	Alum	—								
			6·14-5·36	Silver	107								
15·00-22·00	11·574	0·080-0·107				R180	WG19	51			J, K	0·510 x 0·255	13·0 x 5·8
18·00-26·50	14·047	0·043-0·058	20·7-14·8	Brass	53	R220	WG20	42			J, K	0·420 x 0·170	11·0 x 4·3
			17·6-12·6	Alum	121								
			13·3-9·5	Silver	66								
22·00-33·00	17·328	0·034-0·048				R260	WG21	34			K	0·340 x 0·170	8·6 x 4·3
			–	Brass	–								
			–	Alum	–								
26·50-40·00	21·081	0·022-0·031	21·9-15·0	Silver	96	R320	WG22	28	Q	Ka	K	0·280 x 0·140	7·1 x 3·6
33·00-50·00	26·342	0·014-0·020	–	Brass	—	R400	WG23	22			K, L	0·224 x 0·112	5·7 x 2·9
			31·0-20·9	Silver	97								
40·00-60·00	31·357	0·011-0·015				R500	WG24	19			L	0·188 x 0·094	4·8 x 2·4
50·00-75·00	39·863	0·0063-0·0090	—	Brass	—	R620	WG25	15		V	L, M	0·148 x 0·074	3·8 x 1·9
			52·9-39·1	Silver	98								
60·00-90·00	48·350	0·0042-0·0060	—	Brass	—	R740	WG26	12	O		M	0·122 x 0·061	3·1 x 1·6
			93·3-52·2	Silver	99								
75·00-110·00	59·010	0·0030-0·0041				R900	WG27	10			M	0·100 x 0·050	2·4 x 1·3
90·00-140·00	73·840	0·0018-0·0026	152-99	Silver	138	R1200	WG28	8			M	0·080 x 0·040	2·0 x 1·0
110·0-170·00	90·840	0·0012-0·0017	163-137	Silver	136		WG29	7				0·065 x 0·0325	1·7 x 0·82
140·00-220·00	115·750	0·00071-0·00107	308-193	Silver	135		WG30	5				0·051 x 0·0255	1·3 x 0·65
170·00-260·00	131·520	0·00052-0·00075	384-254	Silver	137		WG31	4				0·043 x 0·0215	1·1 x 0·55
220·00-325·00	173·280	0·00035-0·00047	512-348	Silver	139		WG32	3				0·034 x 0·017	0·87 x 0·44

Rectangular WG Propagation of TE_{10} (H_{10}) Mode



Frequency $f = \frac{\omega}{2\pi} = \frac{c}{\lambda}$ [Hz]

ω – angular velocity [radians/sec]
λ – free space wavelength [m]
c – velocity of light [$\simeq 3 \times 10^8$ m/sec]
a,b, – waveguide dimension as shown on diagram [m]

Wavelength in Guide $\lambda g = \frac{2\pi}{\beta} = \frac{\lambda}{\sqrt{1 - \left(\frac{\lambda}{\lambda o}\right)^2}}$ where $\lambda g > \lambda$

β – phase constant based on phase velocity [radian/sec]
λo – guide cut off wavelength [cm]

Phase velocity $Vp = \frac{\omega}{\beta}$ [cm/sec]

Cut off frequency $fo = \frac{c}{2a}$ [Hz] $c = 3 \times 10^{10}$ [cm/sec]

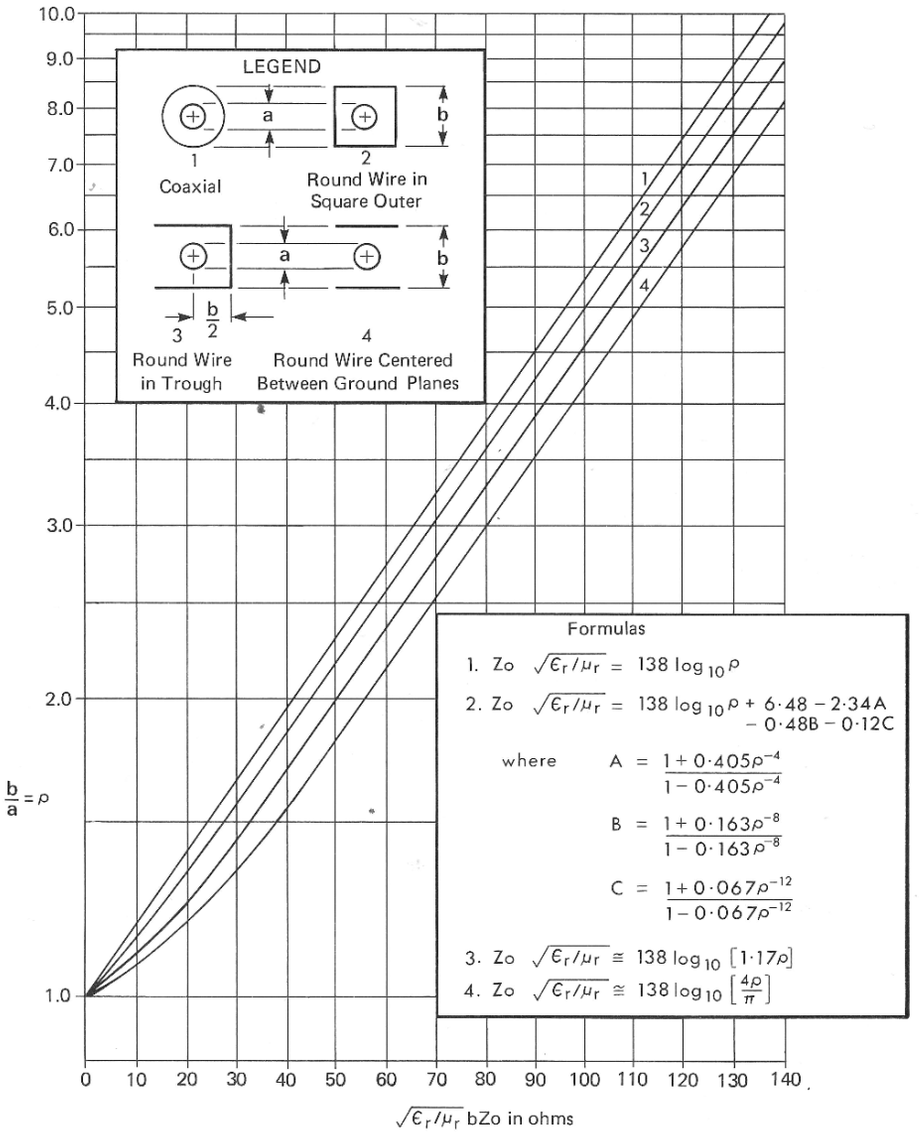
Cut off wavelength $\lambda o = 2a$ [cm]

Phase constant $\beta = \frac{2\pi}{\lambda g} = \frac{2\pi}{\lambda}\sqrt{1 - \left(\frac{\lambda}{\lambda o}\right)^2}$

$= \sqrt{\frac{\omega^2}{c^2} - \frac{\pi^2}{a^2}}$

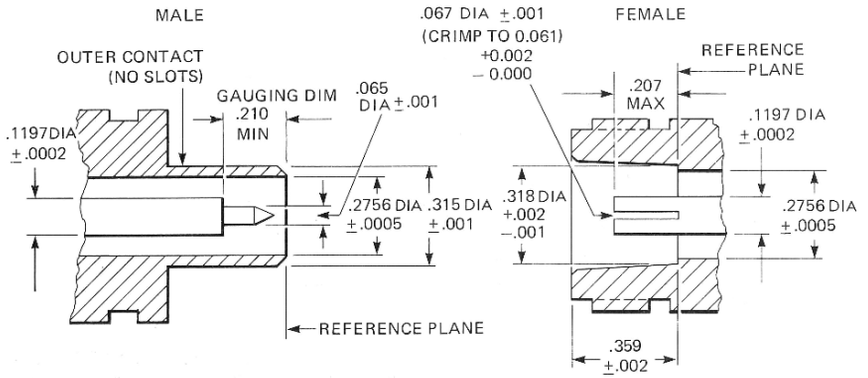
TRANSMISSION LINES

Characteristic Impedance

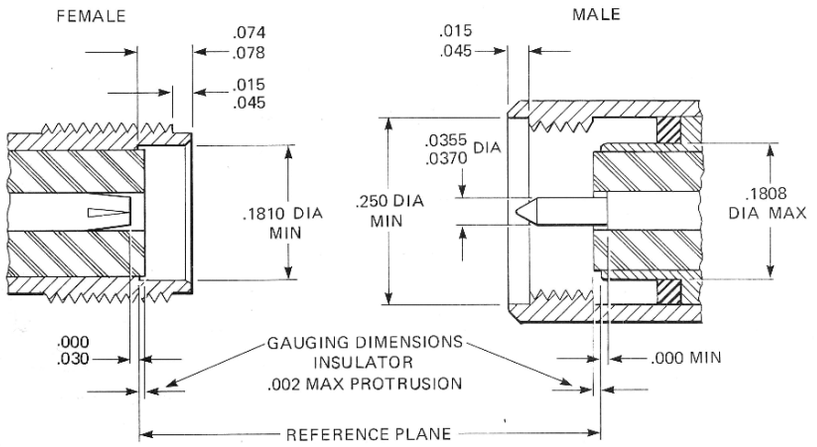


Microwave Coaxial Connector Interface Dimensions

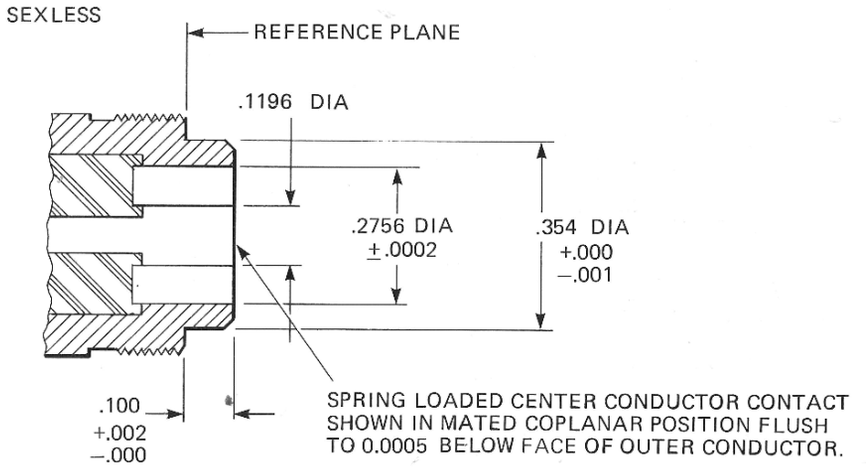
The Key interface dimension of the most general coaxial connectors used at microwave frequencies up to 34 GHz are given, with dimensions in inches.



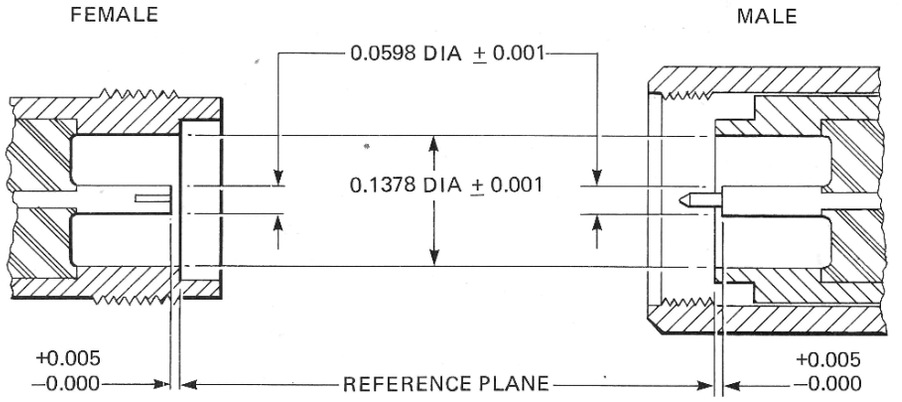
Type N Precision Connector Class II to MIL-C-39012/1 & 2



Type SMA Class II Connectors to MIL-C-39012/55 & 57

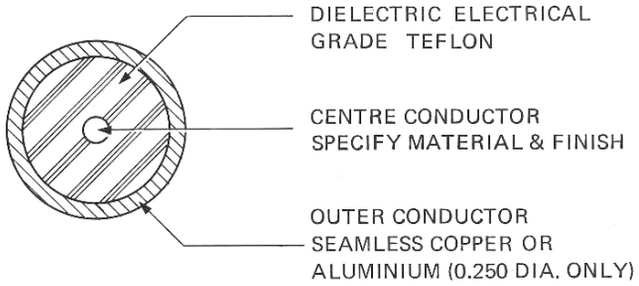


Type APC-7 Precision Connector (IEEE STD No. 287)



Type APC-3-5 Precision Connectors.
Mate with SMA Connector to MIL-C39012/55 & 57

Solid Dielectric Semi Rigid Cable Data


DIELECTRIC ELECTRICAL GRADE TEFLON
CENTRE CONDUCTOR SPECIFY MATERIAL & FINISH
OUTER CONDUCTOR SEAMLESS COPPER OR ALUMINIUM (0.250 DIA. ONLY)

	RG No.	405/U	402/U	401/U	401/U(1)
Mechanical Data	Inner Conductor O.D. (in.)	0·0201±0·001	0·0359±0·0005	0·0641±0·001	0·0641±0·001
	Dielectric O.D. (in.)	0·066±0·003	0·1175±0·001	0·210 ±0·002	0·210 ±0·002
	Outer Conductor O.D. (in.)	0·085±0·001	0·141 ±0·001	0·250 ±0·002	0·250 ±0·002
	Safe Bend Radius (that produced with proper tool) in.(cm)	0·125 (0·32)	0·250 (0·64)	0·375 (0·95)	0·375 (0·95)
	Weight – lbs./100 ft.	1·31	3·25	10·0	6·38
	Max. Operating Temp.(°C)	125	150	175	150
Electrical Characteristic	Impedance Tolerance (ohms)	±1	±1	±1	±1
	Dielectric Strength Rating – KV r.m.s. 60 Hz	5·0	5·0	7·5	7·5
	Corona Extinction Voltage – KV r.m.s. 60 Hz	1·5	1·9	3·15	3·15
	90% Higher Mode Cut-off Frequency – GHz	44·9	30·4	17·2	17·2

A (Attenuation, dB/100 ft at 20°C) and P (Average Power Rating, kW, at unity VSWR in still air at 40°C)

Frequency GHz	A	P	A	P	A	P	A	P
0·10 GHz	5·6	0·785	3·6	2·20	1·9	4·90	2·1	4·6
0·40 GHz	11·6	0·370	7·2	1·00	4·3	2·10	4·6	2·0
1·0 GHz	18·7	0·222	11·6	0·60	7·3	1·20	7·9	1·1
3·0 GHz	34·0	0·115	21·5	0·31	14·0	0·60	14·5	0·57
5·0 GHz	46·0	0·082	28·5	0·23	18·9	0·44	20·4	0·41
10·0 GHz	68·5	0·048	44·5	0·16	29·0	0·28	30·0	0·26
12·0 GHz	84·0	0·040	51·0	0·14	33·0	0·25	34·0	0·23
18·0 GHz	115·0	0·028	68·0	0·10	–	– –	–	–

(1)Aluminium Outer Conductor

Skin Effect and Skin Depth

At high frequencies the current carried by a conductor is not uniformly distributed over the conductor cross section, as at dc, but tends to be concentrated near the surface, where the magnetic flux density is lowest. The thickness within which the current is effectively concentrated is termed the **Skin Depth** given by:

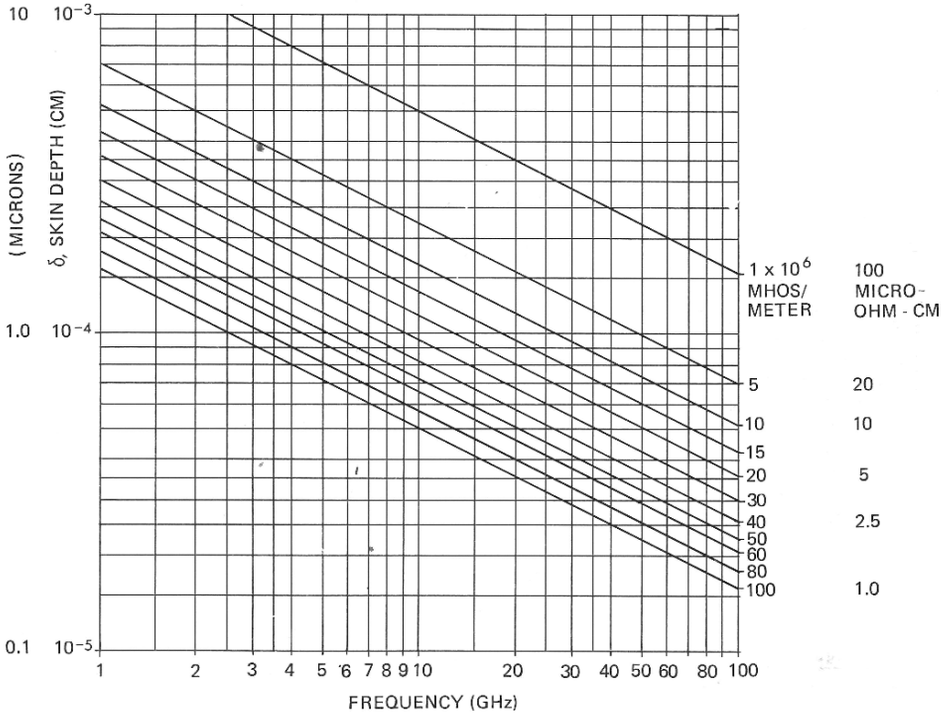
$$\delta = 5033 \sqrt{\frac{\rho}{\mu f}} \quad [\text{cm}]$$

where ρ – resistivity [Ω cm]

μ – magnetic permeability, $\mu = 1$ for non-magnetic materials

f – frequency [Hz]

The graph below shows the skin depth for various conductivities at various frequencies.



Resistivity and conductivity of some common metals:

	microohm-cm	mhos/m		microohm-cm	mhos/m
Silver	1·6	63 x 10^6	Brass 70/30	6·7	15 x 10^6
Copper	1·7	59 x 10^6	Platinum	12	8·3 x 10^6
Gold	2·4	42 x 10^6	Chromium	13	7·7 x 10^6
Aluminium	2·9	34 x 10^6	Tantulum	14	7·1 x 10^6

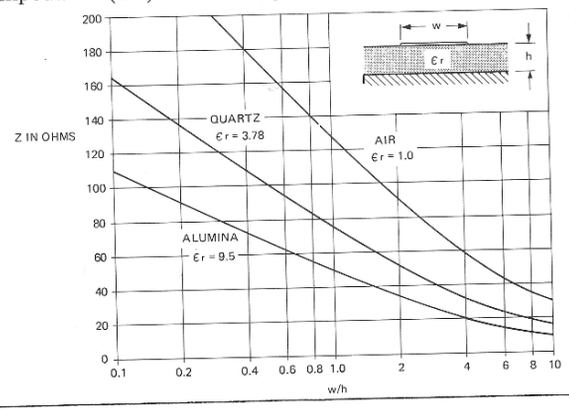
Characteristics of Dielectric Materials

Material	Dielectric Constant	Loss Factor	Useful Temp. Range (°C)	Flexibility	Coeff. of Thermal Expansion ($x10^{-6}$/°C)	Surface Finish (Micro-Inches)
Woven TFG	2·55	0·0015	–60° +200°	Good	18·5	N/A
Microfiber TFG (Duroid 5870)	2·33	0·0005	–60° +200°	Good	5	N/A
Microfiber TFG (Duroid 5880)	2·2	0·0006	–60° +200°	Good	32	N/A
Polystyrene	2·53	0·0003	–60° +100°	Very Poor	7	N/A
Reinforced Polystyrene	2·62	0·002	–60° +100°	Poor	5·7	N/A
Polyphenelene Oxide (PPO)	2·55	0·0016	–60° +200°	Fair	29	N/A
Polyolefin	2·32	0·00015	–60° +100°	Excel-lent	4·4	N/A
Quartz Teflon	2·47	0·0006	–60° +200°	Good	18·5	N/A
Polymide (Micaply 5032)	4·8	0·01	–60° +250°	Good	9	N/A
Epsilam-10	10·0	0·002	–60° +150°	Good	11 to 23	N/A
99·5% Alumina	9·9	0·00008	up to 500°	Very Poor	7·5	<3
Quartz (Fused Silica)	3·78	0·0001	up to 500°	Very Poor	0·55	<1
Sapphire	9·4 11·6	0·00008	up to 500°	Very Poor	7 7 8·3	<1
99·5% BeO	6·6	0·0004	up to 500°	Very Poor	7·8	3 to 7
Boron Nitride	4·4	0·0003	up to 500°	Poor	1 2	N/A

TFG – Teflon Glass Laminates

Microstrip Design

The relationship of track width (w), substrate (thickness 'h', dielectric constant ε_r and characteristic impedance (Zo) are shown graphically.



Transmission Line Theory

Summary of Transmission Line Equations

Quantity	General Line Expression	Ideal Line Expression
Propagation constant	$\gamma = \alpha + j\beta = \sqrt{(R + j\omega L)(G + j\omega C)}$	$\gamma = j\omega\sqrt{LC}$
Phase constant ß	Imaginary part of γ	$\beta = \omega\sqrt{LC} = \frac{2\pi}{\lambda}$
Attenuation constant α	Real part of γ	0
Characteristic impedance	$Z_o = \sqrt{\frac{R + j\omega L}{G + j\omega C}}$	$Z_o = \sqrt{\frac{L}{C}}$
Input impedance	$Z_{-l} = Z_o \frac{Z_r + Z_o \tanh \gamma l}{Z_o + Z_r \tanh \gamma l}$	$Z_{-l} = Z_o \frac{Z_r + jZ_o \tan \beta l}{Z_o + jZ_r \tan \beta l}$
Impedance of short-circuited line. $Z_r = 0$	$Z_{s.c.} = Z_o \tanh \gamma l$	$Z_{s.c.} = jZ_o \tan \beta l$
Impedance of open-circuited line. $Z_o = \infty$	$Z_{o.c.} = Z_o \coth \gamma l$	$Z_{o.c.} = -jZ_o \cot \beta l$
Impedance of line an odd number of quarter wavelengths long	$Z = Z_o \frac{Z_r + Z_o \coth \alpha l}{Z_o + Z_r \coth \alpha l}$	$Z = \frac{Z_o^2}{Z_r}$
Impedance of line an integral number of half wavelengths long	$Z = Z_o \frac{Z_r + Z_o \tanh \alpha l}{Z_o + Z_r \tanh \alpha l}$	$Z = Z_r$
Voltage along line	$V_{-l} = V_i (1 + \Gamma_o e^{-2\gamma l})$	$V_{-l} = V_i (1 + \Gamma_o e^{-2j\beta l})$
Current along line	$I_{-l} = I_i (1 - \Gamma_o e^{-2\gamma l})$	$I_{-l} = I_i (1 - \Gamma_o e^{-2j\beta l})$
Voltage reflection coefficient	$\Gamma = \frac{Z_r - Z_o}{Z_r + Z_o}$	$\Gamma = \frac{Z_r - Z_o}{Z_r + Z_o}$

Useful Approximations for small VSWR's

If u and v are both small with respect to unity,

(a) $(1+u)(1+v) = 1+u+v$

(b) $(1-u)(1-v) = 1-u-v$

(c) $(1 \pm u)^2 = 1 \pm 2u$

(d) $\frac{1+v}{1+u} = 1+v-u$

Examples:

(a) If $r_1 = 1{\cdot}03$ and $r_2 = 1{\cdot}08$, then $r_1 r_2 = 1{\cdot}11$

(b) If $\Gamma_1 = 0{\cdot}96$ and $\Gamma_2 = 0{\cdot}95$, then $\Gamma_1\Gamma_2 = 0{\cdot}91$

(c) If $r = 1{\cdot}04$, $r^2 = 1{\cdot}08$

If $\Gamma = 0{\cdot}96$, $\Gamma^2 = 0{\cdot}92$

(d) If $r_1 = 1{\cdot}03$ and $r_2 = 1{\cdot}08$, then $\frac{r_2}{r_1} = 1{\cdot}05$

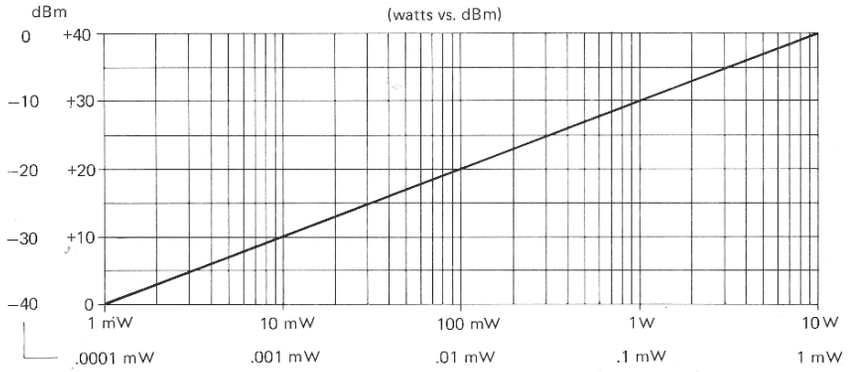
Miscellaneous Relations in Transmission Lines

Equation	Explanation
$r = \frac{1+\lvert\Gamma\rvert}{1-\lvert\Gamma\rvert}$	r = VSWR
$\lvert\Gamma\rvert = \frac{r-1}{r+1}$	$\lvert\Gamma\rvert$ = magnitude of reflection coefficient
$\Gamma = \frac{R-Z_0}{R+Z_0}$	Γ = reflection coefficient (real) at a point in a line where impedance is real (R)
$r = \frac{R}{Z_0}$	$R > Z_0$ (at voltage maximum)
$r = \frac{Z_0}{R}$	$R < Z_0$ (at voltage minimum)
$\frac{P_r}{P_i} = \lvert\Gamma\rvert^2 = \left(\frac{r-1}{r+1}\right)^2$ $\frac{P_t}{P_i} = 1-\lvert\Gamma\rvert^2 = \frac{4r}{(r+1)^2}$	P_r = reflected power P_i = incident power P_t = transmitted power
$\frac{P_b}{P_m} = \frac{1}{r}$	P_b = net power transmitted to load at onset of breakdown in a line where VSWR = r exists. P_m = same when line is matched, r = 1
$\frac{\alpha_r}{\alpha_m} = \frac{1+\lvert\Gamma\rvert}{1-\lvert\Gamma\rvert} = \frac{r^2+1}{2r}$	α_m = attenuation constant when r = 1, matched line α_r = attenuation constant allowing for increased ohmic loss caused by standing waves
$\lvert\Gamma\rvert = \frac{\lvert X\rvert}{\sqrt{\lvert X\rvert^2+4}}$ $\lvert X\rvert = \frac{r-1}{\sqrt{r}}$	Relations for a normalized reactance X in series with resistance Z_0
$\lvert\Gamma\rvert = \frac{\lvert B\rvert}{\sqrt{\lvert B\rvert^2+4}}$ $\lvert B\rvert = \frac{r-1}{\sqrt{r}}$	Relations for a normalized susceptance B in shunt with admittance Y_0

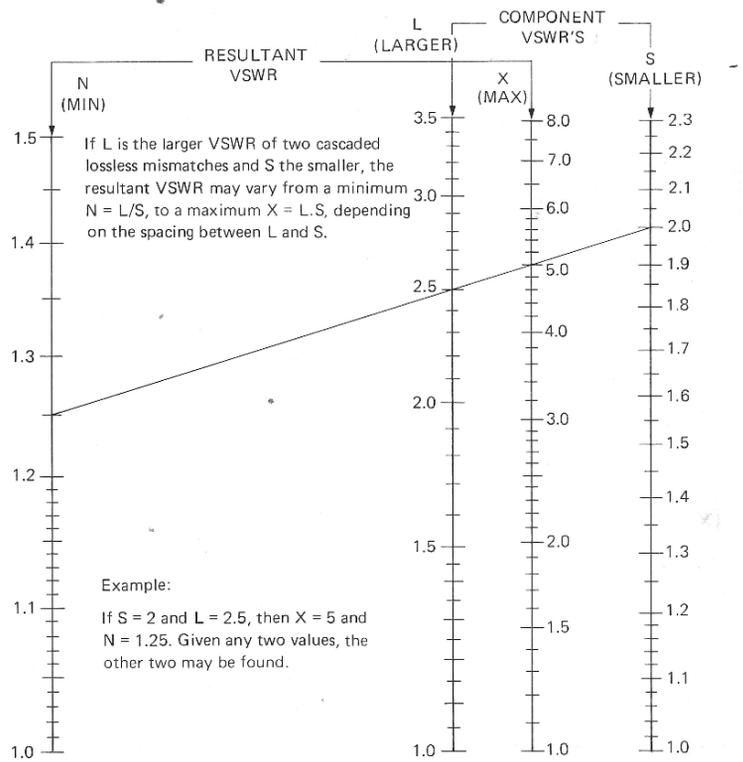
POWER MEASUREMENT

KT = -174 dBm/MHz

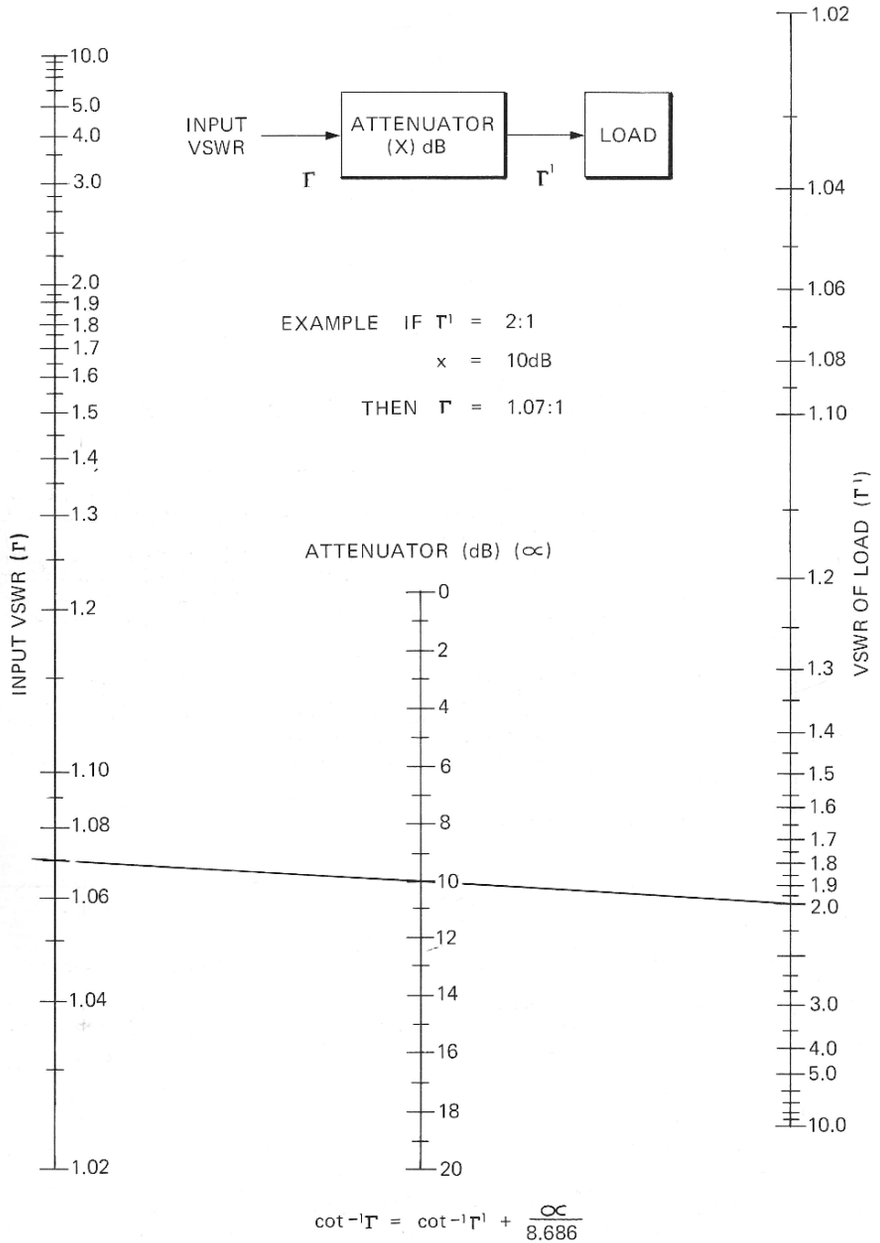
Power Ratio dBm – mW – W.



VSWR Resulting from Two Mismatches

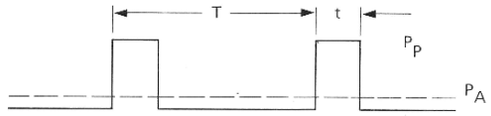

L
(LARGER)
COMPONENT
VSWR'S
RESULTANT
VSWR
S
(SMALLER)
N
(MIN)
X
(MAX)
If L is the larger VSWR of two cascaded lossless mismatches and S the smaller, the resultant VSWR may vary from a minimum N = L/S, to a maximum X = L.S, depending on the spacing between L and S.
Example:
If S = 2 and L = 2.5, then X = 5 and N = 1.25. Given any two values, the other two may be found.

Reduction of VSWR by Matched Attenuators



$$\cot^{-1}\Gamma = \cot^{-1}\Gamma^{1} + \frac{\propto}{8.686}$$

Relationship of Peak Pulse Power to Mean Power



AVG. POWER $P_A = P_P.D.fr$

WHERE P_P – Peak Power

D – Duty cycle = t.fr = $\frac{t}{T}$

fr – repetition frequency

PEAK POWER $P_P = \frac{P_A}{D}$

Measurement of Pulse-risetime Calculations

The risetime of any passive or active transmission network is normally defined as the time of rise of the output voltage, from 10 to 90% of its final value (see Fig. 1) when a perfect step – i.e. one with zero risetime – is applied to the input terminal. The shape of the output waveform is, of course, a function of time that may be expressed as $v/V = f(t)$, where v is the instantaneous voltage and V is the peak voltage.

For practical considerations, however, the expression can be rewritten in the more general form $v_{out} = Av_{in}f_1(t)$, where A is the gain of the network and v_{in} is the input voltage. This expression accommodates the normal case where the input step has a finite rise time, so that v_{in} is also a function of time which can be written as $Vf_2(t)$. And the ultimate waveform is the product of two functions: $v = V.f_1(t).f_2(t)$.

The precise shape of the output step must, of course, depend on the shape of the input step and the h.f. response and phase characteristics of the transmission network. But, for risetime assessment, only the 10% and 90% points on the transient are considered, ignoring effects at upper and lower extremes; and, provided overshoot is not excessive, the transient may be regarded as having Gaussian form. A close approximation to the risetime of the output waveform is obtainable by taking the quadrature sum of the input risetime and the risetime of the network; i.e.,

$$T_{out} = \sqrt{(T^2_{in} + T^2_{network})}$$

where T_{out}, T_{in}, and $T_{network}$ are the respective risetimes.

Composite risetimes

Where two or more networks are connected in cascade, the input voltage to each successive stage would follow the time function of the preceding one if a perfect step were applied to its input. So the risetime of the combination is given by the expression

$$T_{out} = \sqrt{(T_1^2 + T_2^2 + T_3^2 \text{ etc.})}$$

where T_{out} is the risetime of the total, and T_1, T_2, etc. are the risetimes of the individual networks.

The accompanying nomograph gives the value of T_{out} for values of T_1 and T_2. It can also be used when more than two networks are involved by simply regarding each pair of networks as a single network having the combined risetime. For example, if three amplifiers, having risetimes T_1, T_2, and T_3, are connected in cascade, first find $T_x = \sqrt{(T_1^2 + T_2^2)}$, then find $T_{out} = \sqrt{(T_x^2 + T_3^2)}$.

Measuring pulse risetimes

The risetime of an oscilloscope's Y deflection system is, of course, that of the transmission network linking the input terminal to the display screen. In other words, it is the 10 to 90% risetime of the displayed waveform when a perfect step is applied to the Y input.

When an oscilloscope is used to measure pulse risetimes which are long compared with the Y deflection risetime negligible error is introduced, and no correction is necessary. However, when the input risetime is of the same order as the oscilloscope's risetime, that

of the displayed waveform is very nearly equal to their quadrature sum.

Provided the oscilloscope risetime is known, the true risetime of the input waveform can be found from the nomograph by taking T_{out} as the measured risetime, T_1 as the oscilloscope risetime, and T_2 as the risetime of the input waveform.

Network bandwidth

The risetime of an active or passive network having a Gaussian upper-frequency roll-off characteristic is related to the network bandwidth by the expression

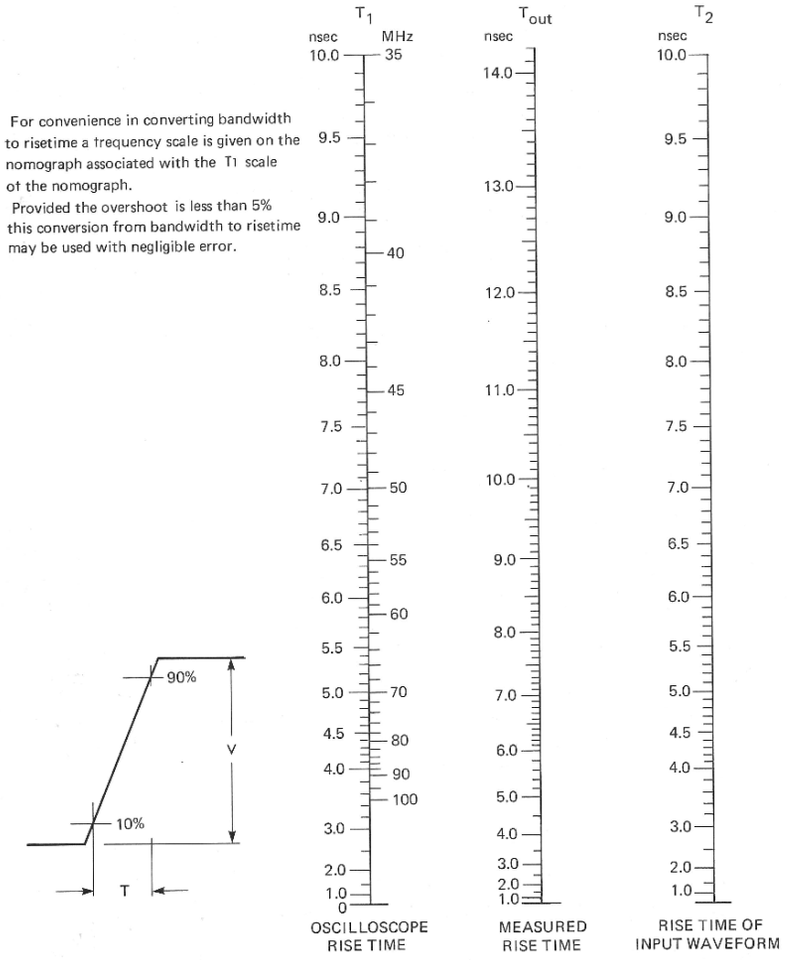
$$f_{-3dB} = 0{\cdot}35/T_{network}$$

where f_{-3dB} is the upper frequency limit at which the response is 3dB below the mid-frequency response, and $T_{network}$ is the network risetime.

NOTE: Any convenient multiplying factor can be applied, in order to bring a practical risetime within the range of the nomograph, provided the same factor is applied to T_1, T_2 and T_{out}. If this factor is 'N' the frequency must be multiplied by $\frac{I}{N}$

For convenience in converting bandwidth to risetime a frequency scale is given on the nomograph associated with the T1 scale of the nomograph.

Provided the overshoot is less than 5% this conversion from bandwidth to risetime may be used with negligible error.



Microwave Power Measurement Uncertainties

When making power measurements at microwave frequencies the following are the most significant sources of uncertainty.

1. **Instrumentation Error** – usually ½ to 1%
   Due to component tolerances in measuring instruments.

2. **Mismatch Uncertainty – The Major Source of Error** – can be ±10%
   In a practical microwave system, with characteristic impedance Z_0 (usually 50 ohms), neither the generator nor the transmission line nor the load (powermeter) will have exactly Z_0 impedance. Furthermore the actual impedances are almost never known completely, being usually only given as VSWR's, which lack phase information. Hence the maximum power is never transferred from the generator to the load and the mismatch loss can only be described as lying between two limits. This uncertainty increases with VSWR and is the fundamental reason for reducing the VSWR of microwave components.

   Two main bases of analysis of mismatch exist. If the power actually delivered to the load is to be compared with the maximum available from the source this is on a **conjugate basis.** For this maximum power transfer the load impedance must be the conjugate of that of the source. This condition may be achieved by using a stub tuner. This is a tedious procedure and hence almost never used in practice.

   If comparison is made with the power delivered by the source to a Z_0 load, then this is on a **Z_0 basis.**

   The basic formula for power transfer between a source and a load (or the fraction of maximum available power absorbed by load) is

$$\frac{(1-|\Gamma s|^2)(1-|\Gamma L|^2)}{|1-\Gamma s\Gamma L|^2}$$

where $|\Gamma|$ is the modulus of the reflection coefficient of source (S) or load (L). This is related to the VSWR (r) by

$$|\Gamma| = \frac{r-1}{r+1}$$

The basic equation takes into account

(a) Z_0 mismatch losses of the source $|\Gamma s|^2$

(b) Z_0 mismatch losses of the load $|\Gamma L|^2$

(c) The relative phase conscious factor $|-\Gamma s\Gamma L|^2$

where Γs, ΓL are complex

If the **conjugate basis** of mismatch analysis is used then all three terms are necessary. With the **Z_0 basis** only the last two terms need be considered.

The Z_0 mismatch loss can have a positive or negative value. With the graph below, the Z_0 mismatch loss and the uncertainty can be deduced.

The Uncertainty of power transfer is

$$-20 \log (1 \pm |\Gamma s|\,|\Gamma L|) \text{ in dB}$$

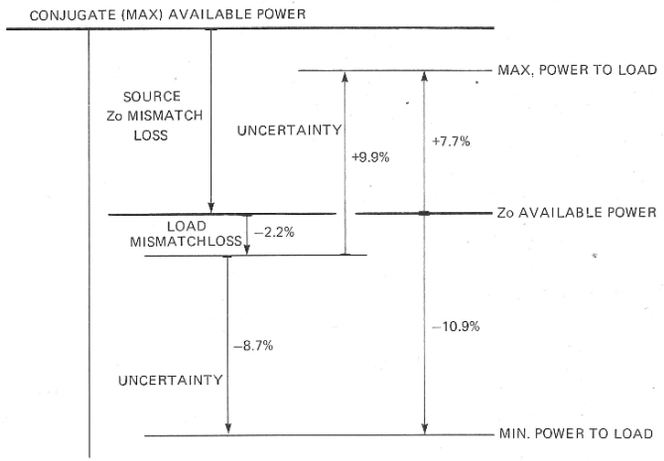
$$\text{or } \{(1 \pm |\Gamma s|\,|\Gamma L|)^2 - 1\} \times 100\%$$

EXAMPLE: With a load VSWR 1·35:1 and source VSWR 1·9:1.

Z_0 Mismatch loss of load = –2·2% (A)

With an added uncertainty of –8·7% to +9·9% (B & C)

The source Z_o mismatch loss is –9·4%



3. **Microwave to low frequency conversion**

Efficiency Variation – usually up to 10% but good correction possible.

The microwave power sensor conversion efficiency and losses within the sensor not measured by the detecting element (usually a thermocouple, bolometer or diode) will vary with frequency.

This can be **corrected** by applying the calibration correction usually provided with power sensor. This calibration can be in the form of either:

CALIBRATION FACTOR – This correction factor allows for:

(a) Microwave to DC substitution error in sensor detector.
(b) RF losses in sensor not measured by sensor detector.
(c) Error due to sensor non-perfect VSWR assuming that power source has perfect VSWR. THIS CONDITION RARELY ACHIEVED. HENCE CALIBRATION FACTOR ALTHOUGH PROVIDING SOME CORRECTION CANNOT PROVIDE PERFECT CORRECTION.

EFFECTIVE EFFICIENCY – This correction factor allows for:

(a) Microwave to DC substitution error in sensor detector.
(b) RF losses in sensor not measured by sensor detector.

4. **Conclusion**

Hence it can be seen that

(1) The Instrumentation error of 1% is relatively small.

(2) The efficiency variation of the power sensor with frequency can be corrected for. The error in this correction is generally a few percent, i.e. the order the uncertainty of the basic power standard.

(3) The mismatch uncertainty (remember that when the load is connected in place of the power sensor, there will be a further mismatch loss and uncertainty) – is however the largest (±10% is not unusual).

How to Improve Microwave Power Measurement Uncertainties

To improve power measurement accuracy the VSWR of the Powermeter AND the Source must be reduced.

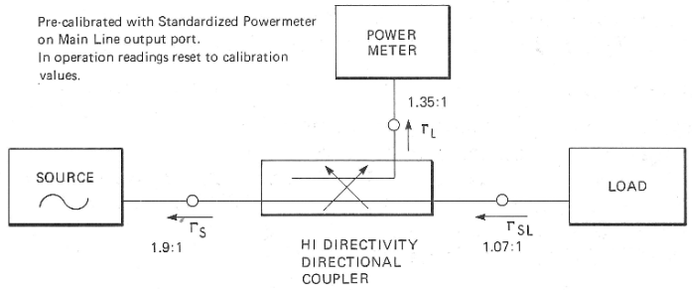
The Powermeter sensor is always designed to provide the lowest possible VSWR. The Source VSWR may be effectively reduced by the following.

(1) Using an **external Attenuator** to effectively mask poor source VSWR (see p.24 for nomograph).

Advantage – simple and economical.
Disadvantage – Insertion loss reduces power to power meter and attentuation measurement uncertainty increases potential error (0·1 dB/10 dB is 2·3%/10 dB).

(2) **Isolator**

Advantage – simple.
Disadvantage – not too economical, narrow band, same uncertainties.



(3) **High Directivity coupler** – will not couple power reflected by load into secondary arm.

Advantage – provides very good equivalent source match.
– low insertion loss.
– power range of system can be extended by coupling factor.
– calibration of coupler/powermeter is independent of source mismatch.

Disadvantage – powermeter operates on lower range which may be less stable.
– additional uncertainty if coupling factor assumed or measured. This can be overcome by calibrating coupler/powermeter combination.

Note

A high directivity directional coupler can provide the equivalent of a low reflection (and if used with stub-tuner then zero reflection). It can be shown [1] that the effective source VSWR (r_{se}) of the system shown is a function only of:

(a) Directional coupler main arm VSWR (typically 1·05).
(b) Coupling factor of directional coupler (say between 10 – 20 dB).
(c) Coupler Directivity (40 dB for WG16).

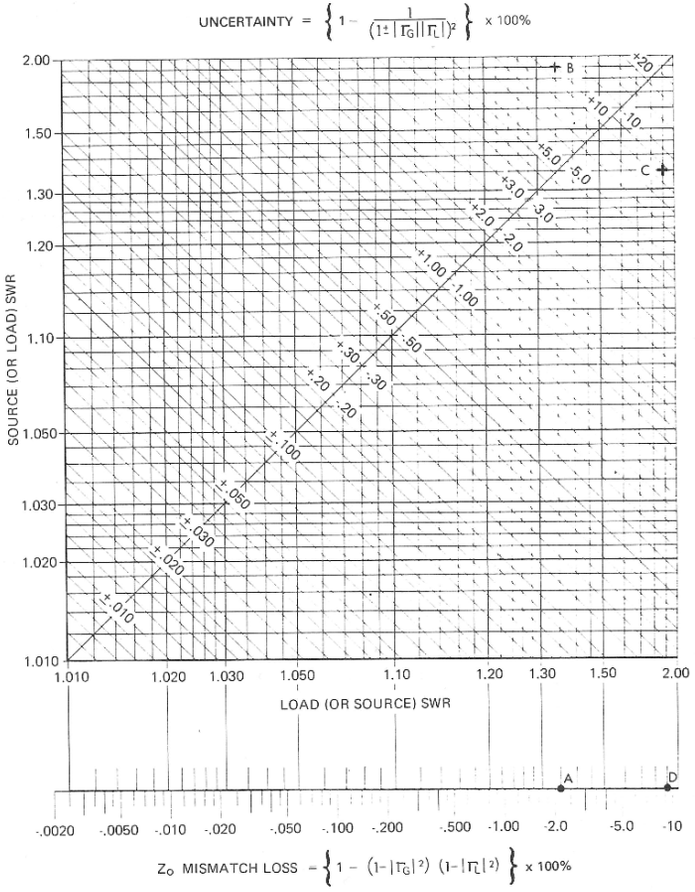
With the above coupler an effective source VSWR of 1·07:1 can be produced.

Example: Using the data from last example, but with source VSWR reduced to 1·07:1, Powermeter VSWR 1·35:1

from graph Z_0 mismatch loss = –2%, but uncertainty now = ±1%.

Hence dramatic improvement in microwave power measurement possible by using high directivity coupler/powermeter combination.

(1) *Rumfelt & Elwell, Radio Frequency Power Measurements, Proc. IEEE, 55, 6, 1967.*



Z_0 mismatch loss and uncertainty chart calibrated in percent of measured power.

Attenuation and Mismatch Terms

1. Substitution Loss:
   A general term referring to the change in power absorbed by a load when first one waveguide junction (these words are used in the broadest sense) and then another (such as attenuator) is used to connect a source to the load.
2. Transducer Loss:
   What the substitution loss becomes when the first waveguide junction is a perfect transducer and initially the maximum available power is delivered to the load.
3. Insertion Loss:
   What the substitution loss becomes when the first waveguide junction is perfect (lossless and phase-shift-less) connector, so that in general there is some initial mismatch loss.
4. Attenuation:
   What any of the above losses becomes when the source and load both have Z_0 impedance. Under these conditions, what is measured on an attenuator is a property of the attenuator alone, so that this is the ideal system in which to make measurements.
5. Residual Attenuation:
   The minimum attenuation of a variable attenuator. What is measured when a variable attenuator is set to its minimum position and inserted into an ideal system.
6. Incremental Attenuation:
   The change in attenuation between minimum setting and any other setting on a variable attenuator. Residual Attenuation and Incremental Attenuation together make up Attenuation.
7. Conjugate Match:
   The condition for maximum power absorption by a load, in which the impedance seen looking toward the load at a point in a transmission line is the complex conjugate of that seen looking toward the source.
8. Conjugate Mismatch:
   The condition in the situation above in which the load impedance is not the conjugate of the source impedance.
9. Conjugate Mismatch Loss:
   The loss resulting from conjugate mismatch.
10. Z_0 Match:
    The condition in which the impedance seen looking into a transmission line is equal to the characteristic impedance of the line.
11. Z_0 Mismatch:
    The condition in which the impedance seen looking into a transmission line is not equal to the transmission line characteristic impedance Z_0. In general, conjugate match is a case of Z_0 mismatch.
12. Z_0 Mismatch Loss:
    The loss resulting from a Z_0 mismatch.
13. Conjugate Available Power:
    Maximum available power.
14. Z_0 Available Power:
    The power a source will deliver to a Z_0 load.

References:
Beatty, R.W., "Intrinsic Attenuation", IEEE Transactions on Microwave Theory and Techniques, Vol. MTT-11, No. 3, May 1963, p. 179. Beatty, R.W. "Insertion Loss Concept", Proceeding IEEE, Vol. 52, No. 6, June 1964, p. 663.

SCALAR NETWORK ANALYSIS

Errors in making transmission and return loss measurements arise from several sources. The largest contribution is usually made by mismatches between elements of the measurement system. Common to all measurements are errors due to:

1. **RF source power variations – with time and load**

   These can be reduced by using a live reference channel and a ratioing technique so that effects of source fluctuations are subtracted from the measurement. Source match errors can be reduced at the expense of reduced dynamic range by using padding attenuators or, over limited bandwidth, by using an isolator.

2. **RF source harmonic content**

   The use of diode detectors which provide a d.c. output related to r.f. input, generally over wide bandwidths, means that any harmonics in the signal will give false readings. The error bounds due to second, third and fourth harmonic is shown related to harmonic level below carrier.

   An r.f. source with a second harmonic level of –20 dBc can give rise to errors of nearly ±1 dB. Hence measurements should be made using r.f. sources with harmonic levels better than –40 dBc.

   A factor which is of special significance in waveguide measurement set-ups is that, if transverse-vane attenuators are included, harmonics will, in general, be passed with far less attenuation than the fundamental.

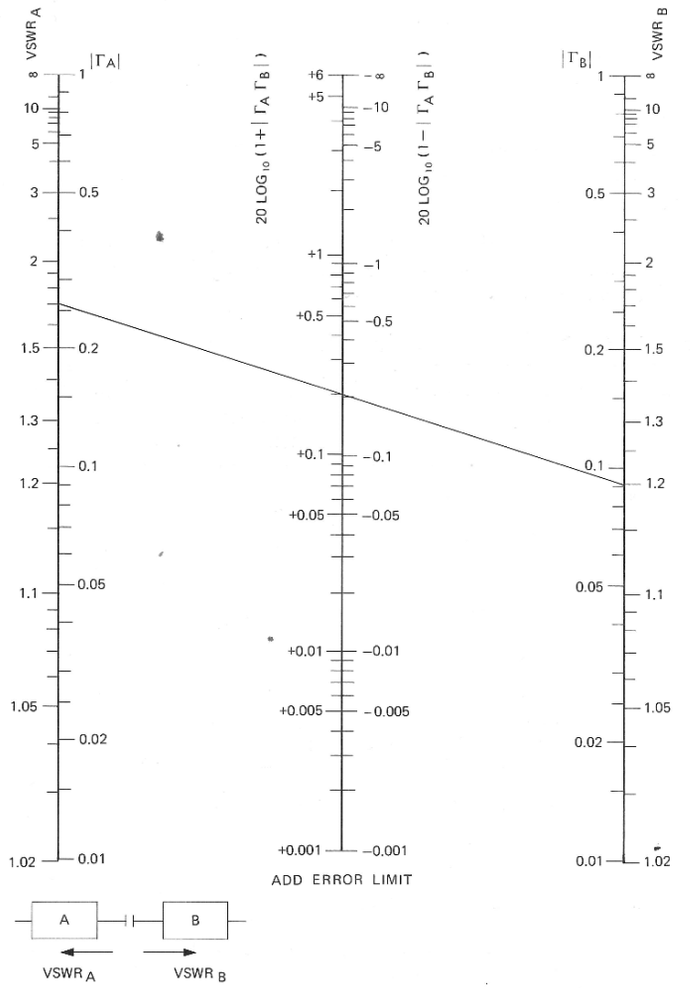
Error effect of RF Source Harmonics

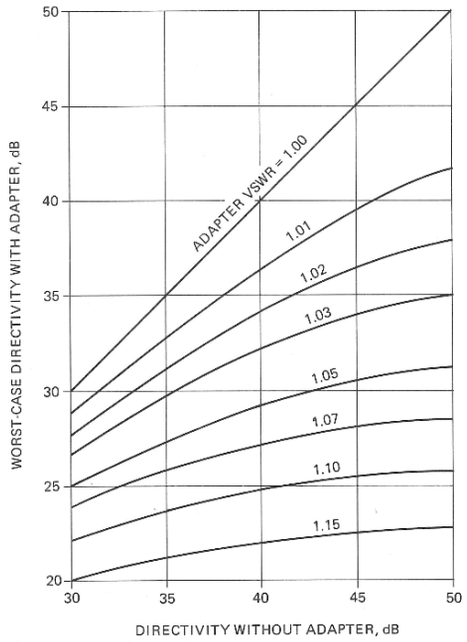
Harmonic Level dBc	Second Harmonic Error Limits (dB)	Third Harmonic Error Limits (dB)	Fourth Harmonic Error Limits (dB)
0	+1·02/+6·02	+3·75/+6·02	+4·65/+6·02
5	–2·11/+3·88	+1·06/+3·88	+2·24/+3·88
10	–2·96/+2·39	–0·62/+2·39	+0·53/+2·39
15	–1·70/+1·42	–1·25/+1·42	–0·40/+1·42
20	–0·92/+0·83	–0·92/+0·83	–0·69/+0·83
25	–0·50/+0·48	–0·50/+0·48	–0·50/+0·48
30	–0·28/+0·27	–0·28/+0·27	–0·28/+0·27
35	–0·16/+0·15	–0·16/+0·15	–0·16/+0·15
40	–0·09/+0·09	–0·09/+0·09	–0·09/+0·09
45	–0·05/+0·05	–0·05/+0·05	–0·05/+0·05
50	–0·03/+0·03	–0·03/+0·03	–0·03/+0·03

Notes:
1. Error = 20 Log Peak (B fundamental + V harmonic) –20 Log (Peak Voltage without harmonics).
2. Total power = P (fundamental) + (harmonic).
   (As measured by thermal effect)

3. **Adaptors**

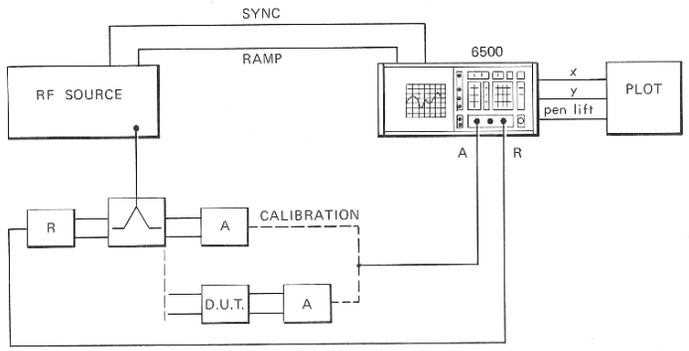
The proliferation of microwave connectors means that in many measurement situations inter-series adaptors are used between test port and items under test. The effect of introducing a further mismatch can be very significant, particularly in return loss measurements as the effective directivity of, say, a bridge with 40 dB directivity could drop to only 28 dB when a 1·05 VSWR adaptor is used at the test port. The effect on Bridge or Coupler directivity is shown on page 34.





Transmission Measurements

The basic measurement set up is shown below using the Marconi Instruments 6500 Automatic Amplitude Analyser.



As well as the errors mentioned earlier, uncertainties arise from power reflections from the mismatch at the interfaces between:

(a) Calibration run; power splitter to detector A.
(b) Measurement run; power splitter to test device.
(c) Measurement run; test device to detector A.

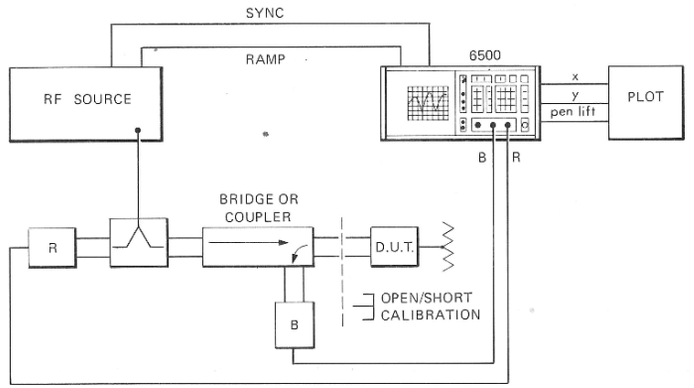
Nomogram on page 33 shows the error limits $20 \log_{10} (1 \pm |\Gamma A \Gamma B|)$ for the two reflection coefficients ΓA and ΓB at the interface. This shows for example, that between two ports of VSWRs 1·7 and 1·2 the possible error is approximately ±0·2 dB in the measured level of the signal passing through that interface. The use of attenuators and isolators to reduce source match problems when measuring low loss items or high VSWR items such as filters is recommended.

In the arrangement shown for Transmission measurements, a power splitter is used to enable source variations to be continuously monitored in the reference channel. Apart from the simplicity of this arrangement, the splitter provides a very good source match at the measurement port since, by virtue of the system ratio-ing action, the input to the splitter is effectively a zero impedance point and both output arms contain precision 50-ohm resistors. Also, splitters are very broadband devices (typically D.C. to 18 GHz), their only drawback being their 6 dB insertion loss. Further, the output VSWR can be measured by conventional methods.

In circumstances in which the insertion loss of a power splitter cannot be tolerated, a directional coupler may be used to provide the reference channel sample. As with the splitter, the ratio-ing process makes the input port of the coupler effectively a zero impedance point, provided that the coupler has high directivity, but this "virtue" presents the user with a dilemma in respect of the output VSWR. It is only the output connector that is of concern because any signal which, having been reflected back into the coupler beyond its output connector and re-reflected at the input end, will become part of the incident signal, sensed by the Reference detector also. The problem is how to measure the VSWR of the output connector alone. If, however, the coupler is *known* to possess high directivity, its input connector must necessarily be almost reflection-free when terminated in a perfect load. So, using a high-grade load, the measured output VSWR will not be greatly in error.

Reflection Measurements

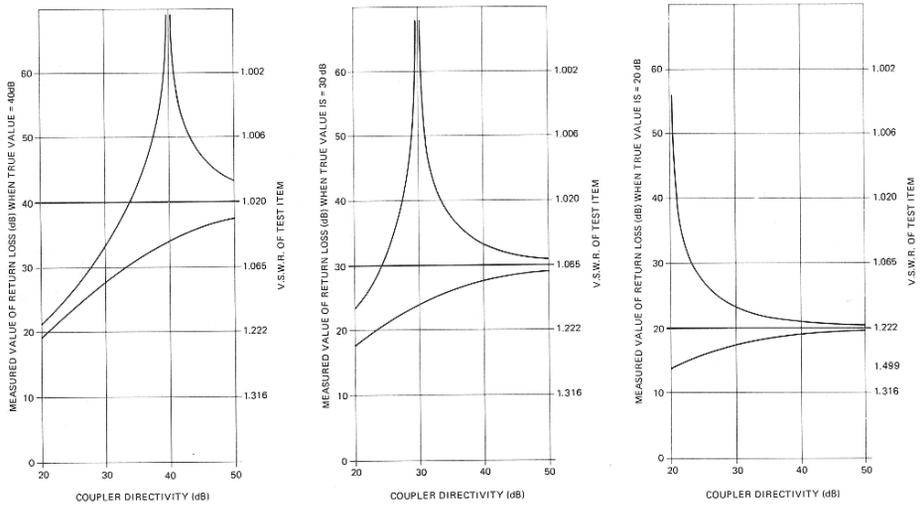
The basic measurement arrangement is shown below.



As well as the problems already mentioned, the measurement of return loss using a bridge or coupler introduces two other errors due to establishing 0 dB return loss for calibration and leakage within the coupler due to low directivity.

The use of an averaged 0 dB reference utilizing open circuit and short circuit terminations will reduce uncertainty since the phasors due to sampling port mismatch will be 180° apart. With 6500, the averaging memory capability can be fully utilized by storing the response with open circuit in memory B and then storing the response of a short circuit averaged with the memory contents.

The Directivity is the ratio of forward to reverse coupling and should be as high as possible since measurement of high return loss (low VSWR) items is subject to very significant errors using low directivity couplers. The use of adaptors as mentioned above degrades directivity. Shown below is the effect introduced by measuring true 20, 30 and 40 dB return loss (1·22, 1·065 and 1·020 VSWR) with couplers of different directivities. For example with a coupler of 40 dB directivity, a measurement on items with a 30 dB return loss could give a reading between 33·6 and 27 dB.



Effect of coupler directivity on return loss measurements.

It is necessary to confirm from time to time, figures for the directivities of couplers used in the measurement system.

The procedure for doing this could be:–

(a) apply a signal to port 1 of the coupler with port 2 terminated and measure the output level at port 3. Coupling Factor equals port 3 level divided by port 1 level.

(b) reverse the coupler, so applying input to port 2 with port 1 terminated, and again measure level at port 3. Directivity equals port 3 level divided by port 2 level, less Coupling Factor, in dB.

However, doubt is cast by the errors due to the return loss of the termination employed. In (a), it is not very significant, but analysis of the uncertainty in (b) is exactly the same as for the uncertainty in using a reflectometer to measure return loss owing to finite coupler directivity. In the graphs above, Directivity and Return Loss may be read interchangeably, depending upon which is being measured.

Programmes can be written to facilitate appreciation of this type of uncertainty calculation.

Conversion of VSWR to Return Loss

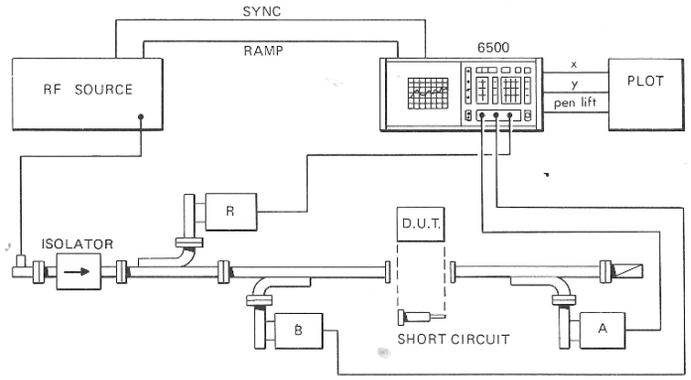
VSWR	SWR (dB)	Return Loss (dB)	Trans. Loss (dB)	Volt. Refl. Coeff.	Power Trans. (%)	Power Refl. (%)
1·00	0·0	∞	0·000	0·00	100·0	0·0
1·01	0·1	46·1	0·000	0·00	100·0	0·0
1·02	0·2	40·1	0·000	0·01	100·0	0·0
1·03	0·3	36·6	0·001	0·01	100·0	0·0
1·04	0·3	34·2	0·002	0·02	100·0	0·0
1·05	0·4	32·3	0·003	0·02	99·9	0·1
1·06	0·5	30·7	0·004	0·03	99·9	0·1
1·07	0·6	29·4	0·005	0·03	99·9	0·1
1·08	0·7	28·3	0·006	0·04	99·9	0·1
1·09	0·7	27·3	0·008	0·04	99·8	0·2
1·10	0·8	26·4	0·010	0·05	99·8	0·2
1·11	0·9	25·7	0·012	0·05	99·7	0·3
1·12	1·0	24·9	0·014	0·06	99·7	0·3
1·13	1·1	24·3	0·016	0·06	99·6	0·4
1·14	1·1	23·7	0·019	0·07	99·6	0·4
1·15	1·2	23·1	0·021	0·07	99·5	0·5
1·16	1·3	22·6	0·024	0·07	99·5	0·5
1·17	1·4	22·1	0·027	0·08	99·4	0·6
1·18	1·4	21·7	0·030	0·08	99·3	0·7
1·19	1·5	21·2	0·033	0·09	99·2	0·8
1·20	1·6	20·8	0·036	0·09	99·2	0·8
1·21	1·7	20·4	0·039	0·10	99·1	0·9
1·22	1·7	20·1	0·043	0·10	99·0	1·0
1·23	1·8	19·7	0·046	0·10	98·9	1·1
1·24	1·9	19·4	0·050	0·11	98·9	1·1
1·25	1·9	19·1	0·054	0·11	98·8	1·2
1·26	2·0	18·8	0·058	0·12	98·7	1·3
1·27	2·1	18·5	0·062	0·12	98·6	1·4
1·28	2·1	18·2	0·066	0·12	98·5	1·5
1·29	2·2	17·9	0·070	0·13	98·4	1·6
1·30	2·3	17·7	0·075	0·13	98·3	1·7
1·32	2·4	17·2	0·083	0·14	98·1	1·9
1·34	2·5	16·8	0·093	0·15	97·9	2·1
1·36	2·7	16·3	0·102	0·15	97·7	2·3
1·38	2·8	15·9	0·112	0·16	97·5	2·5
1·40	2·9	15·6	0·122	0·17	97·2	2·8
1·42	3·0	15·2	0·133	0·17	97·0	3·0
1·44	3·2	14·9	0·144	0·18	96·7	3·3
1·46	3·3	14·6	0·155	0·19	96·5	3·5
1·48	3·4	14·3	0·166	0·19	96·3	3·7
1·50	3·5	14·0	0·177	0·20	96·0	4·0
1·52	3·6	13·7	0·189	0·21	95·7	4·3
1·54	3·8	13·4	0·201	0·21	95·5	4·5
1·56	3·9	13·2	0·213	0·22	95·2	4·8
1·58	4·0	13·0	0·225	0·22	94·9	5·1
1·60	4·1	12·7	0·238	0·23	94·7	5·3
1·62	4·2	12·5	0·250	0·24	94·4	5·6

Conversion of VSWR to Return Loss (Continued)

VSWR	SWR (dB)	Return Loss (dB)	Trans. Loss (dB)	Volt. Refl. Coeff.	Power Trans. (%)	Power Refl. (%)
1·64	4·3	12·3	0·263	0·24	94·1	5·9
1·66	4·4	12·1	0·276	0·25	93·8	6·2
1·68	4·5	11·9	0·289	0·25	93·6	6·4
1·70	4·6	11·7	0·302	0·26	93·3	6·7
1·72	4·7	11·5	0·315	0·26	93·0	7·0
1·74	4·8	11·4	0·329	0·27	92·7	7·3
1·76	4·9	11·2	0·342	0·28	92·4	7·6
1·78	5·0	11·0	0·356	0·28	92·1	7·9
1·80	5·1	10·9	0·370	0·29	91·8	8·2
1·82	5·2	10·7	0·384	0·29	91·5	8·5
1·84	5·3	10·6	0·398	0·30	91·3	8·7
1·86	5·4	10·4	0·412	0·30	91·0	9·0
1·88	5·5	10·3	0·426	0·31	90·7	9·3
1·90	5·6	10·2	0·440	0·31	90·4	9·6
1·92	5·7	10·0	0·454	0·32	90·1	9·9
1·94	5·8	9·9	0·468	0·32	89·8	10·2
1·96	5·8	9·8	0·483	0·32	89·5	10·5
1·98	5·9	9·7	0·497	0·33	89·2	10·8
2·00	6·0	9·5	0·512	0·33	88·9	11·1
2·50	8·0	7·4	0·881	0·43	81·6	18·4
3·00	9·5	6·0	1·249	0·50	75·0	25·0
3·50	10·9	5·1	1·603	0·56	69·1	30·9
4·00	12·0	4·4	1·938	0·60	64·0	36·0
4·50	13·1	3·9	2·255	0·64	59·5	40·5
5·00	14·0	3·5	2·553	0·67	55·6	44·4
5·50	14·8	3·2	2·834	0·69	52·1	47·9
6·00	15·6	2·9	3·100	0·71	49·0	51·0
6·50	16·3	2·7	3·351	0·73	46·2	53·8
7·00	16·9	2·5	3·590	0·75	43·7	56·2
7·50	17·5	2·3	3·817	0·76	41·5	58·5
8·00	18·1	2·2	4·033	0·78	39·5	60·5
8·50	18·6	2·1	4·240	0·79	37·7	62·3
9·00	19·1	1·9	4·437	0·80	36·0	64·0
9·50	19·6	1·8	4·626	0·81	34·5	65·5
10·00	20·0	1·7	4·807	0·82	33·1	66·9
11·00	20·8	1·6	5·149	0·83	30·6	69·4
12·00	21·6	1·5	5·466	0·85	28·4	71·6
13·00	22·3	1·3	5·762	0·86	26·5	73·5
14·00	22·9	1·2	6·040	0·87	24·9	75·1
15·00	23·5	1·2	6·301	0·88	23·4	76·6
16·00	24·1	1·1	6·547	0·88	22·1	77·9
17·00	24·6	1·0	6·780	0·89	21·0	79·0
18·00	25·1	1·0	7·002	0·89	19·9	80·1
19·00	25·6	0·9	7·212	0·90	19·0	81·0
20·00	26·0	0·9	7·413	0·90	18·1	81·9
25·00	28·0	0·7	8·299	0·92	14·8	85·2
30·00	29·5	0·6	9·035	0·94	12·5	87·5

Waveguide Measurements

A typical arrangement is shown below for both transmission and reflection measurements.



The sources of errors are again due to mismatches, source variations and harmonics, coupler directivity etc. Variations between couplers and the use of waveguide to coaxial transitions introduce further uncertainties together with the difficulty of establishing 0 dB return loss calibration, since an open circuit waveguide would radiate. However, the techniques for reducing errors still apply – low source harmonics, use of reference channel, good match between ports. These together with the use of matched, high directivity couplers can provide reasonably accurate transmission and return loss measurements in waveguide systems.

MISCELLANEOUS

IEEE 488-1978 (GPIB) Programming

1. **GPIB Description and Definitions**

The IEEE 488-1978 standard is a byte-serial, bit-parallel interface system primarily designed for use in programmable measurement instrument systems. The general system defines all circuits, cables, connectors, control protocol and message repertoire to ensure the unambiguous transfer of data between devices.

The system has the following defined constraints:

a) No more than 15 devices can be interconnected by a single contiguous bus.
b) Total transmission length is not to exceed 20 meters, or 2 meters times the number of devices, whichever is less.
c) Data rate through any signal line must be less than, or equal to, 1 Megabit/second.
d) All data exchange is to be digital (as opposed to analog).

The IEEE 488-1978 system is structured with 16 transmission lines which consist of:

a) 8 data lines, permitting transfer of ASCII or other binary encoded data. Data transfer is asynchronous and generally bi-directional.
b) 3 data transfer control lines (handshake).
c) 5 general interface management lines.

2. IEEE 488 Capability codes and subsets

The IEEE 488-1978 standard recommends the use of a capability code system to indicate the extent to which each of the interface functions is implemented. On most instruments the capability code will be displayed near the GPIB connector. The following codes are used:

CODE	SUBSET RANGE	INTERFACE FUNCTION
SH	0 – 1	Source Handshake
AH	0 – 1	Acceptor Handshake
T or TE	0 – 8	Talker or Extended Talker
L or LE	0 – 4	Listener or Extended Listener
SR	0 – 1	Service Request
RL	0 – 2	Remote/Local
PP	0 – 2	Parallel Poll
DC	0 – 2	Device Clear
DT	0 – 1	Device Trigger
C	0 – 28	Controller
E	1 – 2	Bus Driver type

In general a zero subset value indicates no capability for the specified function and a non-zero value indicates partial or complete capability. Some functions are inter-dependent and a complete list of allowable subsets is given in IEEE 488-1978 Appendix C.

E1 indicates open-collector bus drivers, which are mandatory for the SRQ, NRFD and NDAC lines.

E2 indicates three-state bus drivers which may be used on all lines except those mentioned above.

3. GPIB Signal Definitions

Data signals:

D101 – D108 DATA INPUT/OUTPUT – used to carry message data in a byte-serial, bit-parallel form. The information carried is interpreted as data or interface commands or device addresses according to the condition of the ATN line.

Handshake signals:

DAV — DATA VALID – used by a talker to indicate the availability and validity of information on the DI0 lines.

NRFD — NOT READY FOR DATA – used to indicate the condition of readiness of listener(s) to accept data.

NDAC — NOT DATA ACCEPTED – used to indicate the condition of acceptance of data by listener(s).

A full description of the handshake timing sequence is given in IEEE 488-1978 Appendix B.

Interface management signals:

ATN — ATTENTION – used by the system controller to indicate how messages on the DI0 lines are to be interpreted.

IFC — INTERFACE CLEAR – used by the system controller to place the interface system (including portions contained within each connected device) into a known quiescent state.

SRQ — SERVICE REQUEST – used by a device to indicate the need for service and to request an interruption of the current sequence of events.

REN — REMOTE ENABLE – used by the system controller in conjunction with other messages to select between two sources of programming data (typically local and remote control).

EOI — END OR IDENTIFY – used by a talker to indicate the end of a data transfer or, in conjunction with the ATN signal, by the system controller to execute a parallel poll sequence.*

The ATN true condition indicates that the DI0 lines are carrying interface commands or device addresses. The interface commands are grouped into ADDRESSED COMMANDS directed at specific devices and UNIVERSAL COMMANDS directed at all devices. The following commands are defined:

Addressed commands:

GTL	(01H)	GO TO LOCAL – instructs a device to return to local control.
SDC	(04H)	SELECTED DEVICE CLEAR – instructs a device to go to a known quiescent state.
*PPC	(05H)	PARALLEL POLL CONFIGURE
GET	(08H)	GROUP EXECUTE TRIGGER – used to trigger a device with DT1 capability.
TCT	(09H)	TAKE CONTROL – used by the system controller to pass control of the interface system to another device with controller capabilities.

Universal commands:

LLO	(11H)	LOCAL LOCKOUT – causes all devices to ignore local control requests. (disables front panel LOCAL keys).
DCL	(14H)	DEVICE CLEAR – instructs all devices to go to a known quiescent state.
*PPU	(15H)	PARALLEL POLL UNCONFIGURE
SPE	(18H)	SERIAL POLL ENABLE – enables devices to participate in serial poll operations.
SPD	(19H)	SERIAL POLL DISABLE – disables serial polling.

*The parallel poll function is not normally supported by Marconi Instruments' products. Instead, the serial poll method of determining the source of, and reason for, a SRQ interrupt is used. With this method each device is polled in a sequence determined by the system controller and returns a status byte on the D10 lines. D107 is used to indicate SRQ active and DI01 – DI06 and DI08 are used to carry device-dependent status information.

4. The Connector for IEEE BUS

Both the dimensions and the actual pin locations of the connector are prescribed in the 488-1975 Standard. Recommended connectors include MICRORIBBON (Amphenol or Cinch Series 57), or CHAMP (AMP).

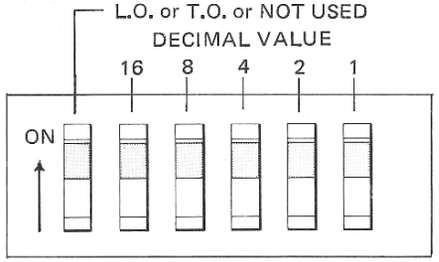
Connector Pin Assignments

Contact	Signal Line	Contact	Signal Line
1	DI01	13	D105
2	DI02	14	DI06
3	DI03	15	DI07
4	DI04	16	DI08
5	EOI	17	REN
6	DAV	18	Gnd. (6)
7	NRFD	19	Gnd. (7)
8	NDAC	20	Gnd. (8)
9	IFC	21	Gnd. (9)
10	SRQ	22	Gnd. (10)
11	ATN	23	Gnd. (11)
12	SHIELD	24	Gnd. LOGIC

Note: Gnd. (n) refers to the signal ground return of the referenced contact.

5. **IEEE 488 and IEC 625 BUS Standards**
These are essentially the same, except for the connector.
The IEC 625 specifies a Cannon 'D' type connector, while the IEEE 488 specifies the MICRORIBBON connector. A preference seems to exist for the IEEE System and most manufacturers including Marconi Instruments have opted for this alternative. An adaptor between the IEEE and IEC System is however available from Marconi Instruments (P/N 46883-408K).

6. **GPIB Address selection**
The valid address range is 0 – 31. In some instruments the GPIB address is selectable by a front panel key-press sequence, but in most cases it is selected by a five position dual-in-line switch on the rear panel. The combination of the five switch positions can give any decimal value between 0 and 31 according to a binary code. A 6th. switch position is usually provided which may enable a TALK ONLY or LISTEN ONLY function.



MI GPIB Address Switch connection

7. **Message data formats**
The IEEE 488-1978 standard defines all the mechanical, electrical and data transfer protocols associated with the interface system. It does not offer detailed recommendations for device-dependent data formats. The more recently published IEEE 728-1982 gives full recommendations for various groups of message data formats. It is the intention that all Marconi Instruments' products will comply in general with IEEE 728-1982.

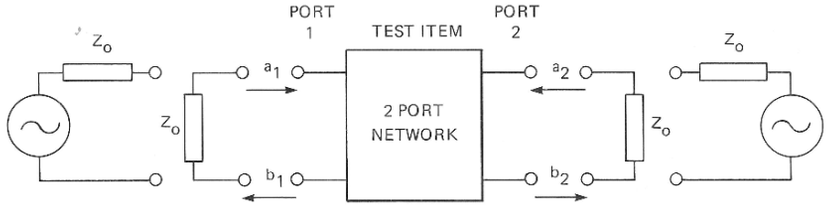
S-Parameters and Transformations

In microwave circuit design S-parameters are very useful for the full characterization of any 2 port Network.

In contrast to z, y and h-parameters, which require broadband short circuited and open circuited connections at the TEST ITEM for the measurement, S-parameters are determined with input and output terminated with the resistive characteristic impedance of test systems (generally 50 ohms in coaxial line system).

Parasitic oscillations in active devices are minimised when these devices are terminated in resistive loads.

S-parameters are complex, having a magnitude and a phase relationship, and are measured in terms of incident and reflected voltages using a VECTOR VOLTMETER.



The four S-parameters are:

With Generator connected to port 1 and port 2 perfectly matched ($a_2 = o$)

INPUT-REFLECTION COEFFICIENT $S_{11} = \frac{b_1}{a_1}$

Looking into port 1 when port 2 is perfectly matched.

FORWARD-TRANSMISSION COEFFICIENT $S_{21} = \frac{b_2}{a_1}$

Voltage transmission coefficient from port 1 to port 2 when port 2 is perfectly matched.

With Generator connected to port 2 and port 1 perfectly matched ($a_1 = o$)

REVERSE-TRANSMISSION COEFFICIENT $S_{12} = \frac{b_1}{a_2}$

Voltage transmission coefficient from port 2 to port 1 when port 1 is perfectly matched.

OUTPUT-REFLECTION COEFFICIENT $S_{22} = \frac{b_2}{a_2}$

Looking into port 2 when port 1 is perfectly matched.

Useful Scattering Parameters Relationships

$$b_1 = s_{11}a_1 + s_{12}a_2$$
$$b_2 = s_{21}a_1 + s_{22}a_2$$

Input reflection coefficient with arbitrary Z_L

$$s'_{11} = s_{11} + \frac{s_{12}s_{21}\Gamma_L}{1 - s_{22}\Gamma_S}$$

Output reflection coefficient with arbitrary Zs

$$s'_{22} = s_{22} + \frac{s_{12}s_{21}\Gamma_S}{1 - s_{11}\Gamma_S}$$

Voltage gain with arbitrary Z_L and Zs

$$A_V = \frac{V_2}{V_1} = \frac{s_{21}(1 + \Gamma_L)}{(1 - s_{22}\Gamma_L)(1 + s'_{11})}$$

$$\text{Power Gain} = \frac{\text{Power delivered to load}}{\text{Power input to network}}$$

$$G = \frac{|s_{21}|^2 (1 - |\Gamma_L|^2)}{(1 - |s_{11}|^2) + |\Gamma_L|^2 (|s_{22}|^2 - |D|^2) - 2\,\mathrm{Re}\,(\Gamma_L N)}$$

$$D = S_{11}S_{22} - S_{12}S_{21}$$
$$M = S_{11} - DS^*_{22}$$
$$N = S_{22} - DS^*_{11}$$

$$\text{Available Power Gain} = \frac{\text{Power available from network}}{\text{Power available from source}}$$

$$G_A = \frac{|s_{21}|^2 (1 - |\Gamma_S|^2)}{(1 - |s_{22}|^2) + |\Gamma_S|^2 (|s_{11}|^2 - |D|^2) - 2\,\mathrm{Re}\,(\Gamma_S M)}$$

$$\text{Transducer Power Gain} = \frac{\text{Power delivered to load}}{\text{Power available from source}}$$

$$G_T = \frac{|s_{21}|^2 (1 - |\Gamma_S|^2)(1 - |\Gamma_L|^2)}{|(1 - s_{11}\Gamma_S)(1 - s_{22}\Gamma_L) - s_{12}s_{21}\Gamma_L\Gamma_S|^2}$$

Unilateral Transducer Power Gain ($s_{12} = 0$)

$$G_{TU} = \frac{|s_{21}|^2 (1 - |\Gamma_S|^2)(1 - |\Gamma_L|^2)}{|1 - s_{11}\Gamma_S|^2 |1 - s_{22}\Gamma_L|^2}$$
$$= G_0 G_1 G_2$$
$$G_0 = |s_{21}|^2$$
$$G_1 = \frac{1 - |\Gamma_S|^2}{|1 - s_{11}\Gamma_S|^2}$$
$$G_2 = \frac{1 - |\Gamma_L|^2}{|1 - s_{22}\Gamma_L|^2}$$

Maximum Unilateral Transducer Power Gain when $|s_{11}| < 1$ and $|s_{22}| < 1$

$$G_U = \frac{|s_{21}|^2}{|(1 - |s_{11}|^2)(1 - |s_{22}|)^2|}$$
$$= G_0 G_{1\,\max} G_{2\,\max}$$
$$G_{i\,\max} = \frac{1}{1 - |s_{ii}|^2} \qquad i = 1, 2$$

This maximum attained for $\Gamma_S = s^*_{11}$ and $\Gamma_L = s^*_{22}$

Constant Gain circles (Unilateral case: $s_{12} = 0$)

– centre of constant gain circle is on line between centre of Smith Chart and point representing s^*_{ii}

– distance of centre of circle from centre of Smith Chart:

$$r_i = \frac{g_i |s_{ii}|}{1 - |s_{ii}|^2 (1 - g_i)}$$

– radius of circle:

$$\rho_i = \frac{\sqrt{1 - g_i}\,(1 - |s_{ii}|^2)}{1 - |s_{ii}|^2 (1 - g_i)}$$

Where: i = 1, 2

$$\text{and } g_i = \frac{G_i}{G_{i\,\max}} = G_i (1 - |s_{ii}|^2)$$

Unilateral Figure of Merit

$$u = \frac{|s_{11}s_{22}s_{12}s_{21}|}{|(1 - |s_{11}|^2)(1 - |s_{22}|^2)|}$$

Error Limits on Unilateral Gain Calculation

$$\frac{1}{(1 + u^2)} < \frac{G_T}{G_{TU}} < \frac{1}{(1 - u^2)}$$

Conditions for Absolute Stability
No passive source or load will cause network to oscillate if a, b, and c are all satisfied.

a. $|S_{11}| < 1, |S_{22}| < 1$

b. $\left| \dfrac{|S_{12}S_{21}| - |M^*|}{|S_{11}|^2 - |D|^2} \right| > 1$

c. $\left| \dfrac{|S_{12}S_{21}| - |N^*|}{|S_{22}|^2 - |D|^2} \right| > 1$

Condition that a two-port network can be simultaneously matched with a positve real source and load:

$K > 1$ or $C < 1$

C = Linvill C factor $= K^{-1}$

$D = S_{11}S_{22} - S_{12}S_{21}$

$M = S_{11} - DS^*_{22}$

$N = S_{22} - DS^*_{11}$

$$K = \frac{1 + |D|^2 - |S_{11}|^2 - |S_{22}|^2}{2|S_{12}S_{21}|} = \text{Rollett Stability Factor}$$

Source and Load for Simultaneous Match

$$\Gamma_{ms} = M^* \left| \frac{B_1 \pm \sqrt{B_1^2 - 4|M|^2}}{2|M|^2} \right|$$

$$\Gamma_{mL} = N^* \left| \frac{B_2 \pm \sqrt{B_2^2 - 4[N]^2}}{2|N|^2} \right|$$

Where $B_1 = 1 + |S_{11}|^2 - |S_{22}|^2 - |D|^2$

$B_2 = 1 + |S_{22}|^2 - |S_{11}|^2 - |D|^2$

Maximum Available Power Gain, MAG

If $K > 1$,

$$MAG = \left| \frac{S_{21}}{S_{12}} (K \pm \sqrt{K^2 - 1}) \right|$$

(Use plus sign when B_1 is positive, minus sign when B_1 is negative. For definition of B_1 see 'Source and Load for Simultaneous Match', above.)

Maximum Stable Gain, MSG

$$MSG = \left| \frac{S_{21}}{S_{12}} \right|$$

Unilateral Gain – Mason

$$U = \frac{1/2\,|(S_{21}/S_{12}) - 1|^2}{K\,|S_{21}/S_{12}| - \text{Re}(S_{21}/S_{12})}$$

s-parameters in terms of h-, y-, and z-parameters	h-, y-, and z-parameters in terms of s-parameters
$S_{11} = \dfrac{(Z_{11} - 1)(Z_{22} + 1) - Z_{12}Z_{21}}{(Z_{11} + 1)(Z_{22} + 1) - Z_{12}Z_{21}}$	$Z_{11} = \dfrac{(1 + S_{11})(1 - S_{22}) + S_{12}S_{21}}{(1 - S_{11})(1 - S_{22}) - S_{12}S_{21}}$
$S_{12} = \dfrac{2Z_{12}}{(Z_{11} + 1)(Z_{22} + 1) - Z_{12}Z_{21}}$	$Z_{12} = \dfrac{2S_{12}}{(1 - S_{11})(1 - S_{22}) - S_{12}S_{21}}$
$S_{21} = \dfrac{2Z_{21}}{(Z_{11} + 1)(Z_{22} + 1) - Z_{12}Z_{21}}$	$Z_{21} = \dfrac{2S_{21}}{(1 - S_{11})(1 - S_{22}) - S_{12}S_{21}}$
$S_{22} = \dfrac{(Z_{11} + 1)(Z_{22} - 1) - Z_{12}Z_{21}}{(Z_{11} + 1)(Z_{22} + 1) - Z_{12}Z_{21}}$	$Z_{22} = \dfrac{(1 + S_{22})(1 - S_{11}) + S_{12}S_{21}}{(1 - S_{11})(1 - S_{22}) - S_{12}S_{21}}$
$S_{11} = \dfrac{(1 - Y_{11})(1 + Y_{22}) + Y_{12}Y_{21}}{(1 + Y_{11})(1 + Y_{22}) - Y_{12}Y_{21}}$	$Y_{11} = \dfrac{(1 + S_{22})(1 - S_{11}) + S_{12}S_{21}}{(1 + S_{11})(1 + S_{22}) - S_{12}S_{21}}$
$S_{12} = \dfrac{-2Y_{12}}{(1 + Y_{11})(1 + Y_{22}) - Y_{12}Y_{21}}$	$Y_{12} = \dfrac{-2S_{12}}{(1 + S_{11})(1 + S_{22}) - S_{12}S_{21}}$
$S_{21} = \dfrac{-2Y_{21}}{(1 + Y_{11})(1 + Y_{22}) - Y_{12}Y_{21}}$	$Y_{21} = \dfrac{-2S_{21}}{(1 + S_{11})(1 + S_{22}) - S_{12}S_{21}}$
$S_{22} = \dfrac{(1 + Y_{11})(1 - Y_{22}) + Y_{12}Y_{21}}{(1 + Y_{11})(1 + Y_{22}) - Y_{12}Y_{21}}$	$Y_{22} = \dfrac{(1 + S_{11})(1 - S_{22}) + S_{12}S_{21}}{(1 + S_{22})(1 + S_{11}) - S_{12}S_{21}}$
$S_{11} = \dfrac{(h_{11} - 1)(h_{22} + 1) - h_{12}h_{21}}{(h_{11} + 1)(h_{22} + 1) - h_{12}h_{21}}$	$h_{11} = \dfrac{(1 + S_{11})(1 + S_{22}) - S_{12}S_{21}}{(1 - S_{11})(1 + S_{22}) + S_{12}S_{21}}$
$S_{12} = \dfrac{2h_{12}}{(h_{11} + 1)(h_{22} + 1) - h_{12}h_{21}}$	$h_{12} = \dfrac{2S_{12}}{(1 - S_{11})(1 + S_{22}) + S_{12}S_{21}}$
$S_{21} = \dfrac{-2h_{21}}{(h_{11} + 1)(h_{22} + 1) - h_{12}h_{21}}$	$h_{21} = \dfrac{-2S_{21}}{(1 - S_{11})(1 + S_{22}) + S_{12}S_{21}}$
$S_{22} = \dfrac{(1 + h_{11})(1 - h_{22}) + h_{12}h_{21}}{(h_{11} + 1)(h_{22} + 1) - h_{12}h_{21}}$	$h_{22} = \dfrac{(1 - S_{22})(1 - S_{11}) - S_{12}S_{21}}{(1 - S_{11})(1 + S_{22}) + S_{12}S_{21}}$

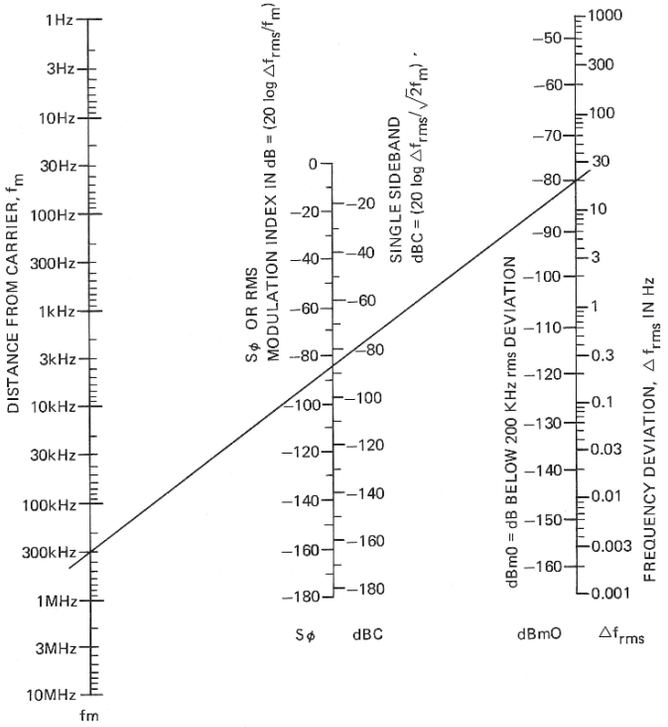
Transistor Frequency Parameters

f_t = frequency at which $|h_{fe}|$
= $|h^{21}$ for common emitter configuration$| = 1$

f_{max} = frequency at which MAG = 1

Noise Conversion Nomograph

Relationship between Modulating frequency (fm), Power spectral density of phase (Sϕ), Modulation Index, Sideband to carrier ratio (dBc), dBm0, and frequency deviation (frms).



Example:
20 Hz deviation in a 1 kHz band at 300 kHz from carrier gives single sideband dBC of –87 dB in a 1 kHz Bandwidth.

Intermodulation Products

Because no amplifier is perfectly linear, intermodulation products are generated when two or more frequencies are amplified simultaneously. Similarly mixers generate intermodulation products.

As a typical example if two signals at frequencies f_1 and f_2 are present within passband of amplifier then 2nd order intermodulation products from 2nd harmonic or direct mixing of the two signals occur. These products appear at 2 f_1, 2 f_2, f_1+f_2 and f_1-f_2.

The 3rd order products result from mixing between two signals f_1 and f_2 and the 2nd order products. These 3rd order products appear at frequencies 2 f_1-f_2, 2 f_1+f_2, 2 f_2-f_1, 2 f_2+f_1. The signals of interest are 2 f_1-f_2 and 2 f_2-f_1 which lie directly in passband.

In most practical cases the two signals f_1 and f_2 are non-equal and a nomograph is used to determine spurious signal level.

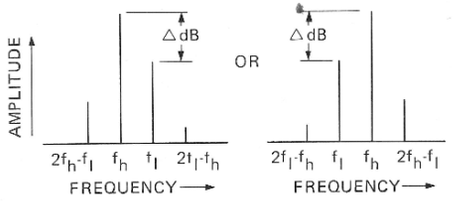
Ex. Amplifier with following characteristics.

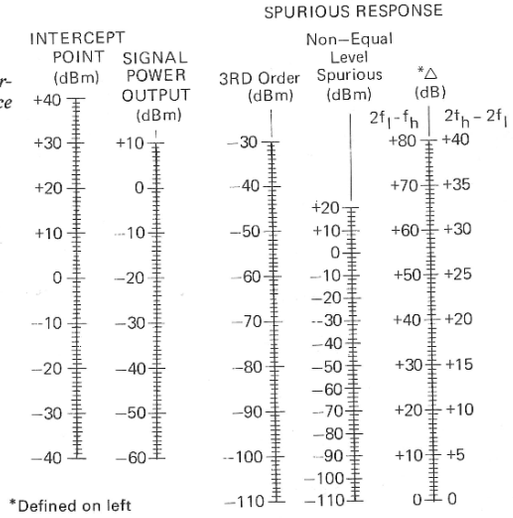
3rd order Intercept Point(1) = 20 dBm
Output Signal at 2 GHz = –10 dBm
and Output Signal at 2·1 GHz = –15 dBm

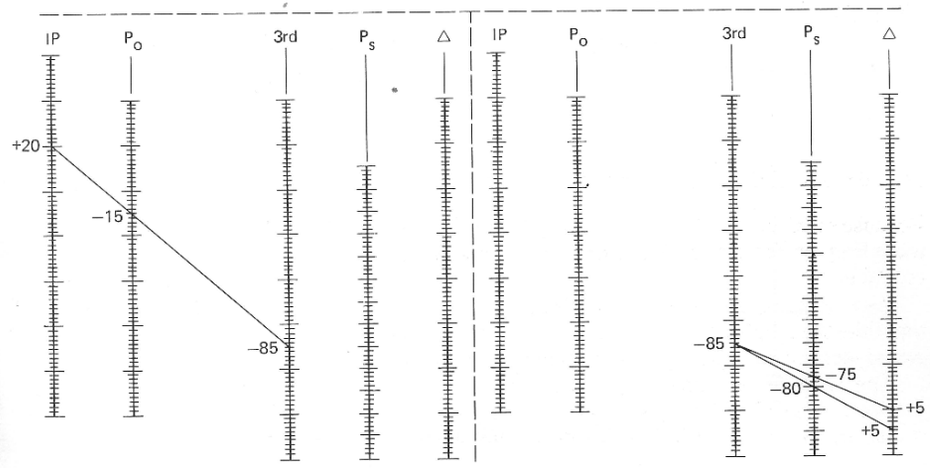
What is amplitude of nearest spurious response?

Choosing lower signal level nomograph shows 3rd order spurious response of device is –85 dBm. For 2 signals $\Delta = 5$ dB amplitude difference nomograph shows one spurious signal is –75 dBm at 1·9 GHz and other –80 dBm at 2·2 GHz.

(1) *Intercept Point (dBm) is a figure of merit for intermodulation product suppression and specified by device manufacturer.*







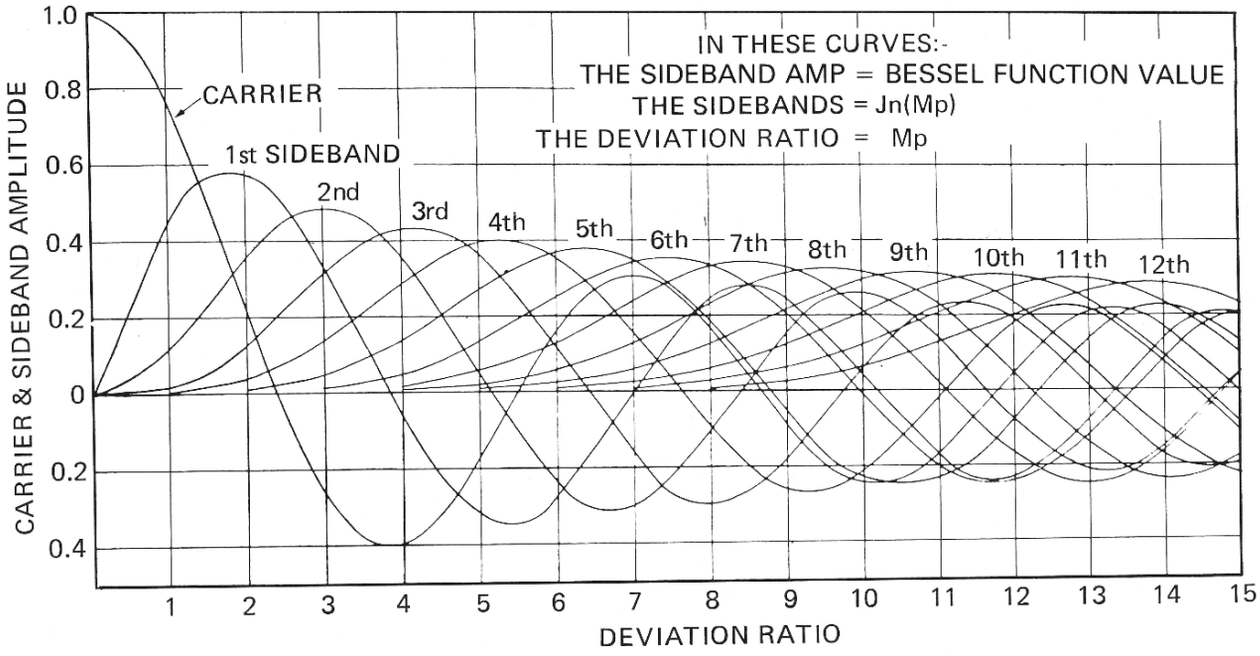


TABLE 1

Bessel function values for deviation ratios less than unity

n	$J_n(0{\cdot}1)$	$J_n(0{\cdot}2)$	$J_n(0{\cdot}3)$	$J_n(0{\cdot}4)$	$J_n(0{\cdot}5)$	$J_n(0{\cdot}6)$	$J_n(0{\cdot}7)$	$J_n(0{\cdot}8)$	$J_n(0{\cdot}9)$	$J_n(1{\cdot}0)$
0	0·9975	0·9900	0·9776	0·9604	0·9385	0·9120	0·8812	0·8463	0·8075	0·7652
1	0·0499	0·0995	0·1483	0·1960	0·2423	0·2867	0·3290	0·3688	0·4059	0·4401
2	—	—	0·0112	0·0197	0·0306	0·0437	0·0588	0·0758	0·0946	0·1149
3	—	—	—	—	—	—	—	0·0102	0·0144	0·0196

NOTE.—Only those values greater than 0·0100 are given.

TABLE 2

Bessel function values for deviation ratios up to 15

n	$J_n(1)$	$J_n(2)$	$J_n(3)$	$J_n(4)$	$J_n(5)$	$J_n(6)$	$J_n(7)$	$J_n(8)$	$J_n(9)$	$J_n(10)$	$J_n(11)$	$J_n(12)$	$J_n(13)$	$J_n(14)$	$J_n(15)$
0	·7652	·2239	−·2601	−·3971	−·1776	·1506	·3001	·1717	−·0903	−·2459	−·1712	·0477	·2069	·1711	−·0142
1	·4401	·5767	·3391	−·0660	−·3276	−·2767	−·0047	·2346	·2453	·0435	−·1768	−·2234	−·0703	·1334	·2051
2	·1149	·3528	·4861	·3641	·0466	−·2429	·3014	−·1130	·1448	·2546	·1390	−·0849	−·2177	−·1520	·0416
3	·0196	·1289	·3091	·4302	·3648	·1148	−·1676	−·2911	−·1809	·0584	·2273	·1951	·0033	−·1768	−·1940
4	—	·0340	·1320	·2811	·3912	·3576	·1578	−·1054	−·2655	−·2196	−·0150	·1825	·2193	·0762	−·1192
5	—	—	·0430	·1321	·2611	·3621	·3479	·1858	−·0550	−·2341	−·2383	−·0735	·1316	·2204	·1305
6	—	—	·0114	·0491	·1310	·2458	·3392	·3376	·2043	−·0145	−·2016	−·2437	−·1180	·0812	·2061
7	—	—	—	·0152	·0534	·1296	·2336	·3206	·3275	·2167	·0184	−·1703	−·2406	−·1508	·0345
8	—	—	—	—	·0184	·0565	·1280	·2235	·3051	·3179	·2250	·0451	−·1410	−·2320	−·1740
9	—	—	—	—	—	·0212	·0589	·1263	·2149	·2919	·3089	·2304	·0670	−·1143	−·2200
10	—	—	—	—	—	—	·0235	·0608	·1247	·2075	·2804	·3005	·2338	·0850	−·0901
11	—	—	—	—	—	—	—	·0256	·0622	·1231	·2010	·2704	·2927	·2357	·0999
12	—	—	—	—	—	—	—	—	·0274	·0634	·1216	·1953	·2615	·2855	·2367
13	—	—	—	—	—	—	—	—	·0108	·0290	·0643	·1201	·1901	·2536	·2787
14	—	—	—	—	—	—	—	—	—	·0119	·0304	·0650	·1188	·1855	·2464
15	—	—	—	—	—	—	—	—	—	—	·0130	·0316	·0656	·1174	·1813
16	—	—	—	—	—	—	—	—	—	—	—	·0140	·0327	·0661	·1162
17	—	—	—	—	—	—	—	—	—	—	—	—	·0149	·0337	·0665
18	—	—	—	—	—	—	—	—	—	—	—	—	—	·0158	·0346
19	—	—	—	—	—	—	—	—	—	—	—	—	—	—	·0166

NOTE.—Only those values greater than 0·0100 are given.

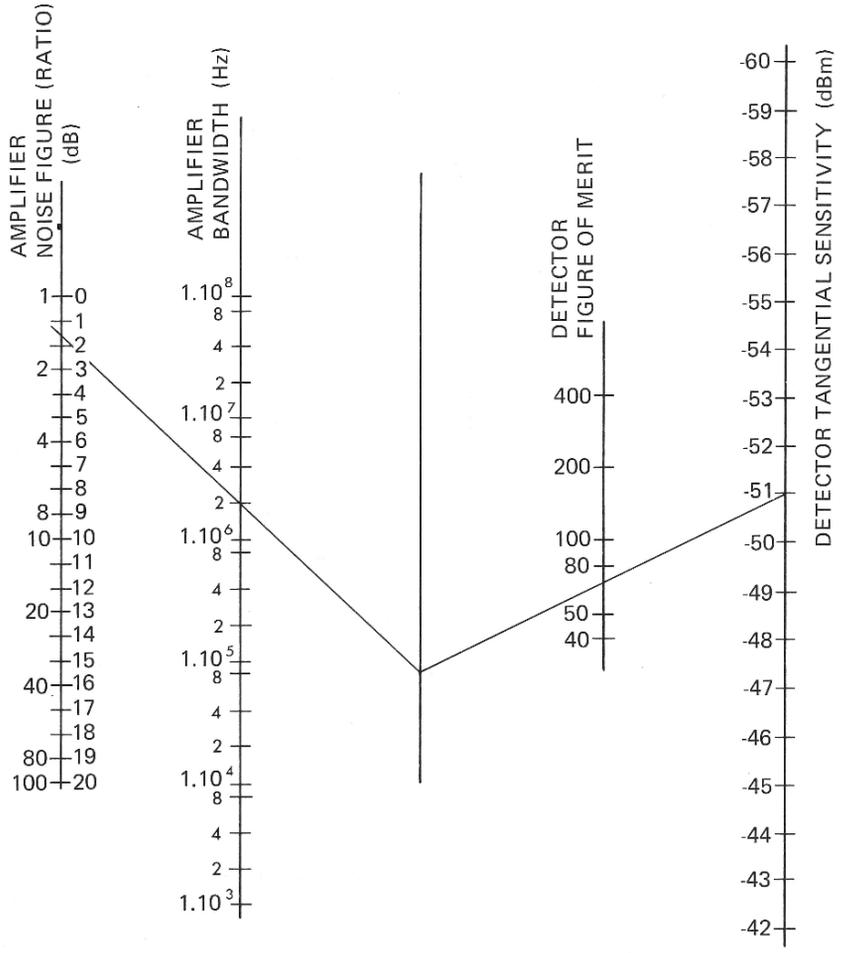
Tangential Sensitivity of a Microwave Detector

To determine the tangential sensitivity of a detector having a known video resistance R_v and open circuit voltage sensitivity K, operating into a video amplifier of known Bandwidth and Noise Figure, first:

Calculate the Figure of Merit (M)

$$M = \frac{K}{\sqrt{R_v}} \qquad K = \frac{\text{millivolts}}{\text{milliwatt}} \qquad R_v = \text{Ohms}$$

Then mark the amplifier noise figure and bandwidth on the appropriate lines on the nomogram below and proceed.

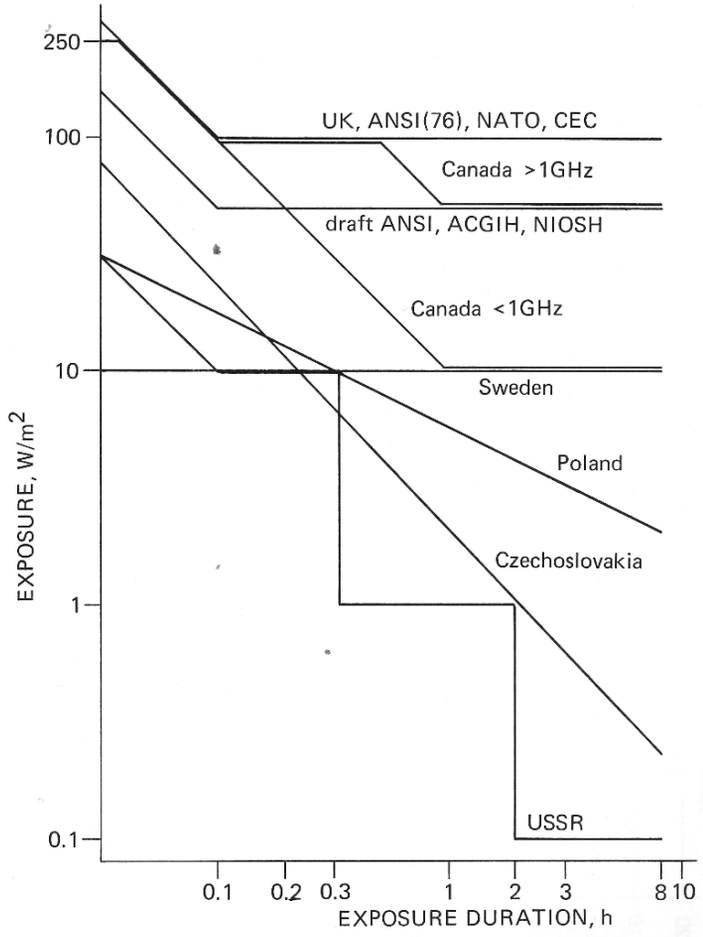


Example: Back Diode Detector

$R_v = 100$ ohms $\qquad K = 700 \ \frac{mV}{mW} \qquad M = \frac{K}{\sqrt{R_v}} = \frac{700}{10} = 70$

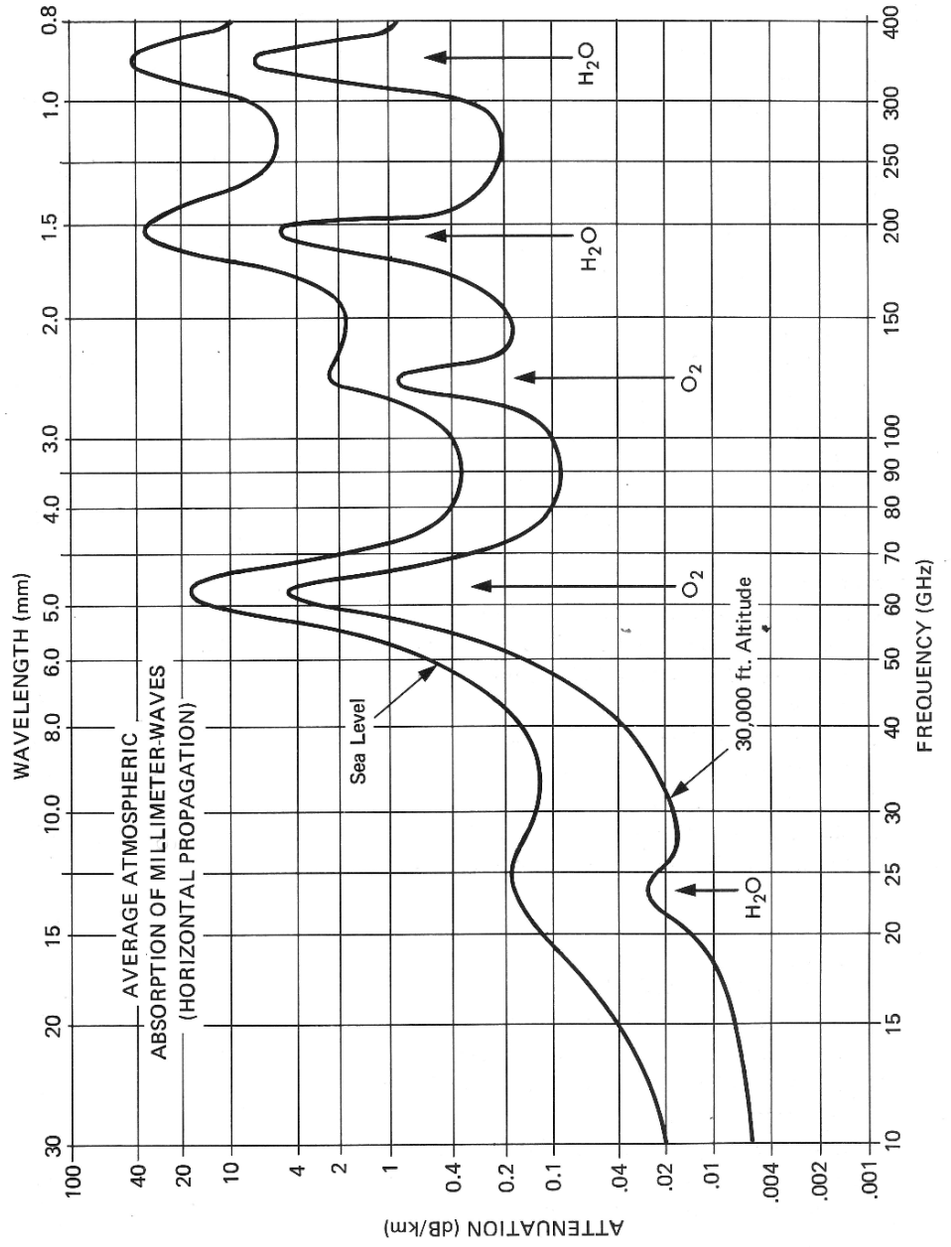
If we follow this detector with a video amp. having 2 MHz bandwidth and 1 ·5 dB N.F., then TSS (determined from the nomogram) is –51 dBm.

Microwave Radiation Hazard Safety Standards



Comparison of whole-body exposure standards for non-ionizing radiation.

Average Atmospheric Absorption of Millimetre Waves



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